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Practical Circuits for Experimenters

W1FB'S DESIGN NOTEBOOK



By Doug DeMaw, W1FB



PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE



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By Doug DeMaw, W1FB



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Foreword

Do you like to build Amateur Radio equipment? Would you like to? If your answer to either of these questions is yes, then this book is for you.

We are pleased to present the fourth volume in the Doug DeMaw notebook series. Two of the other volumes in this series deal with antennas. The third is *QRP Notebook*, and this notebook is related in that it has more of the simple equipment that can be found in the earlier work. In these pages you will find basic radio projects for the bands below 30 MHz. You will also find explanations of how various circuits work. One thing you will not find is heavy mathematical analysis.

This plain-language book is filled with simple, practical projects that can be built using common hand tools. You'll not need exotic or hard-to-find components to build the projects. Nor will you need elaborate test equipment to make them work.

Sound good? We think so too. Please use the feedback form at the back to let us know what you think of this book, and what you'd like to see in future League publications.

David Sumner, K1ZZ
Executive Vice President

Newington, Connecticut
July 1990

Schematic Symbols Used in this Book

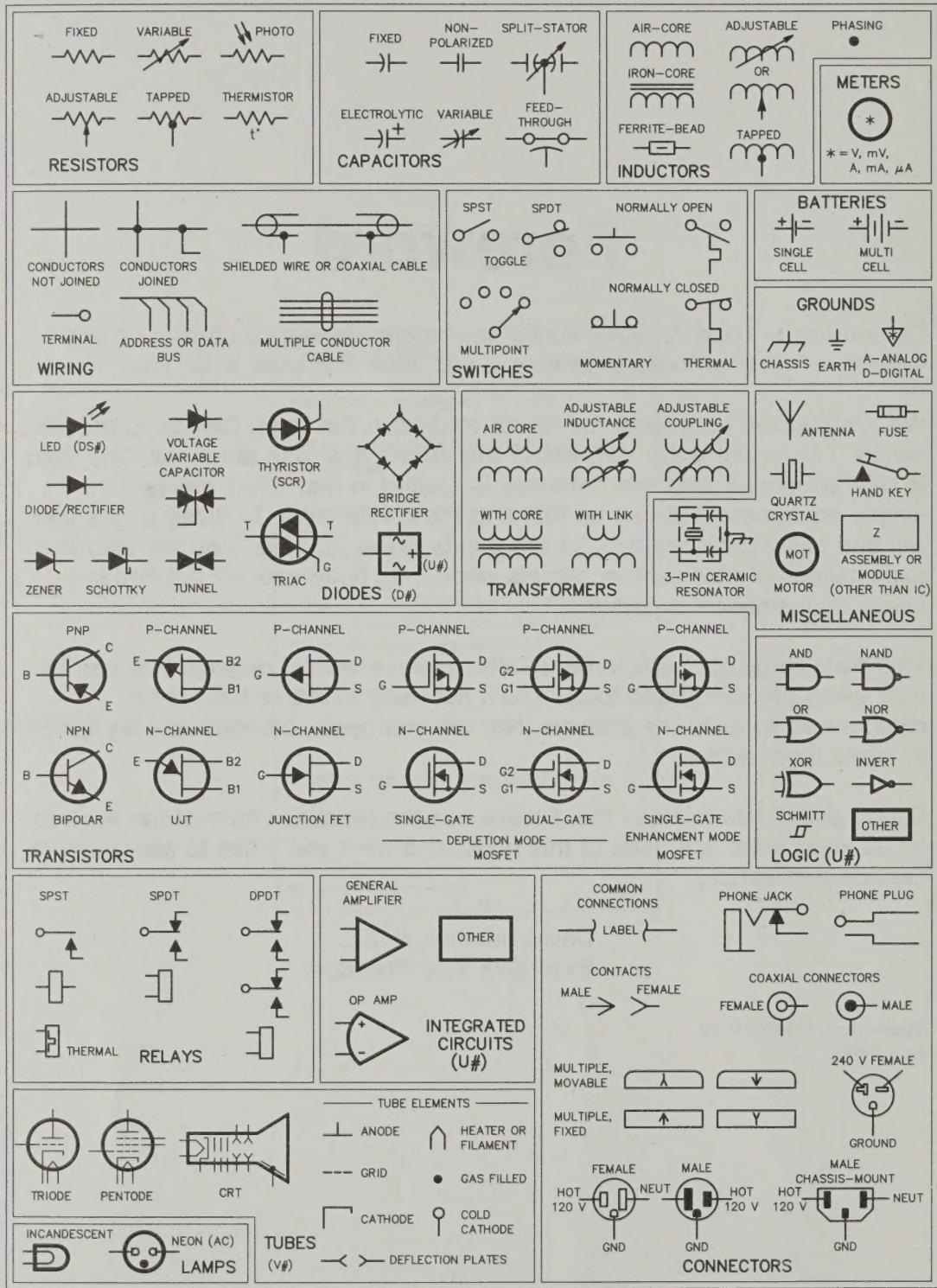


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INTRODUCTION

Amateur Radio is more than a hobby. It is an art that you can enjoy if you are willing to experiment with circuits. Learning to build some of your station equipment can be exciting, and it does not require that you have a formal background in electronics. At the beginning of ham radio it was necessary for all amateurs to build station equipment, however crude it may have appeared after it was completed. Commercial amateur equipment was not available when radio was in its infancy. Ham experimenters built batteries that were used to power their low-power transmitters. Many of the parts were made by hand, such as capacitors, because manufactured units were scarce and expensive. Coils were often hand wound on cardboard forms, such as oatmeal boxes. Shellac was used to hold the coil turns in place. Those were exciting times, because it was a feat to build a transmitter or receiver that worked well. Pride was associated with Amateur Radio as a result of successful experiments and contributions by hams to the state of the art.

You need not be an electrical wizard to build equipment that works well. Doctors, farmers, factory workers, housewives, school-age children and truck drivers build electronic equipment. The process is called "learn by doing." The more you experiment the better your understanding of electronics. This new knowledge can be aided through regular study of **The ARRL Handbook** and some of the more basic ARRL books. Simple projects represent the proper starting place for those with no prior exposure to circuit building. Your first project may be something as simple as a one-transistor mic preamplifier, or perhaps a one-transistor CW transmitter.

Printed-circuit boards are not essential for today's experimenters. You may use point-to-point wiring (called "ugly construction") on a blank piece of PC or perforated board. Learning how to etch PC boards can come later. Your finished project need not be a work of art as long as it performs correctly. There are few amateurs who can impart the "commercial look" to their projects, but pride is experienced because of the circuit performance.

Your workshop does not need to be equipped like an engineering lab in order for you to be successful as an experimenter. Many hams enjoy success with only a volt-ohm-milliammeter and a few piece of home-made test gear. Don't let a lack of test equipment keep you from enjoying the thrills of experimentation.

This book is dedicated to the nontechnical ham who wants to build simple projects and obtain a basic understanding of amateur electronics. Getting involved with circuits will greatly enhance your enjoyment of Amateur Radio.

DIODES, ICs & TRANSISTORS

CHAPTER 1

It is important that we understand the fundamental principles of semiconductors if we are to be successful with our experiments and equipment building. Active devices are those that require an operating voltage in order to function. Transistors, ICs and vacuum tubes are among the active devices we amateurs use. On the other hand, passive devices require no operating voltage to make them work. Examples of passive components are diodes, capacitors, resistors, crystal filters and inductors.

Transistors, unlike vacuum tubes, amplify current. Vacuum tubes amplify voltage. ICs (integrated circuits) and LSIs (large scale integrated circuits) contain transistors, resistors and diodes. Therefore, they contain passive and active components. An IC may, in some instances, contain hundreds of BJTs (bipolar junction transistor), some diodes and many resistors. These are all formed at one time on a single piece of silicon crystal (substrate).

TYPES of TRANSISTORS

You will hear about planar, point-contact, mesa and other varieties of transistor. These terms relate to the manner in which the inner workings of the transistor are formed during the manufacturing process. You will become aware also of the terms FET (field-effect transistor) and MOSFET (metal oxide FET). The FET is a junction field-effect transistor, whereas the MOSFET is an insulated-gate FET. A junction type of FET has its gate, drain and source elements formed into a sandwich, as is the case with bipolar transistors and junction diodes. The MOSFET, on the other hand, has a drain-source junction or sandwich, but the gate or gates (some have two gates) are isolated from this junction by means of a thin layer of oxide. These gates are very sensitive to static charges, which can quickly puncture the insulation and cause a short circuit between the gate and the drain-source junction. MOSFETs must be handled with great care to prevent damage. Similar care must be exercised when soldering a MOSFET into a PC board.

Transistors are manufactured for many operating power levels. Some are made only for use from dc through the audio spectrum, while others are suitable up to and including the microwave spectrum. Transistors may be used not only for amplifying ac and RF energy, but also as electronic switches and dc amplifiers. We can purchase power MOSFETs that work well as RF amplifiers from VLF through VHF. We will discuss these devices in greater detail, later in this chapter.

You will observe that transistors come in a variety of packages or cases. Some are enclosed in metal, while others have plastic cases. Each case style has an assigned industrial designator, such as T0-3, T0-220 or T0-92. This means that each transistor style requires a particular mounting technique. Also, some of these transistors need to be mounted on a heat sink, and the heat sink is chosen (mass) in accordance with the power dissipated by the transistor. Small-

signal transistors (low power) do not require heat sinks in most circuits. Fig 1-1 shows how transistors and vacuum tubes are configured. Note that bipolar transistors and triode tubes each have three similar elements. Dual-gate MOSFETs and tetrode tubes are also similar with regard to the number of device elements.

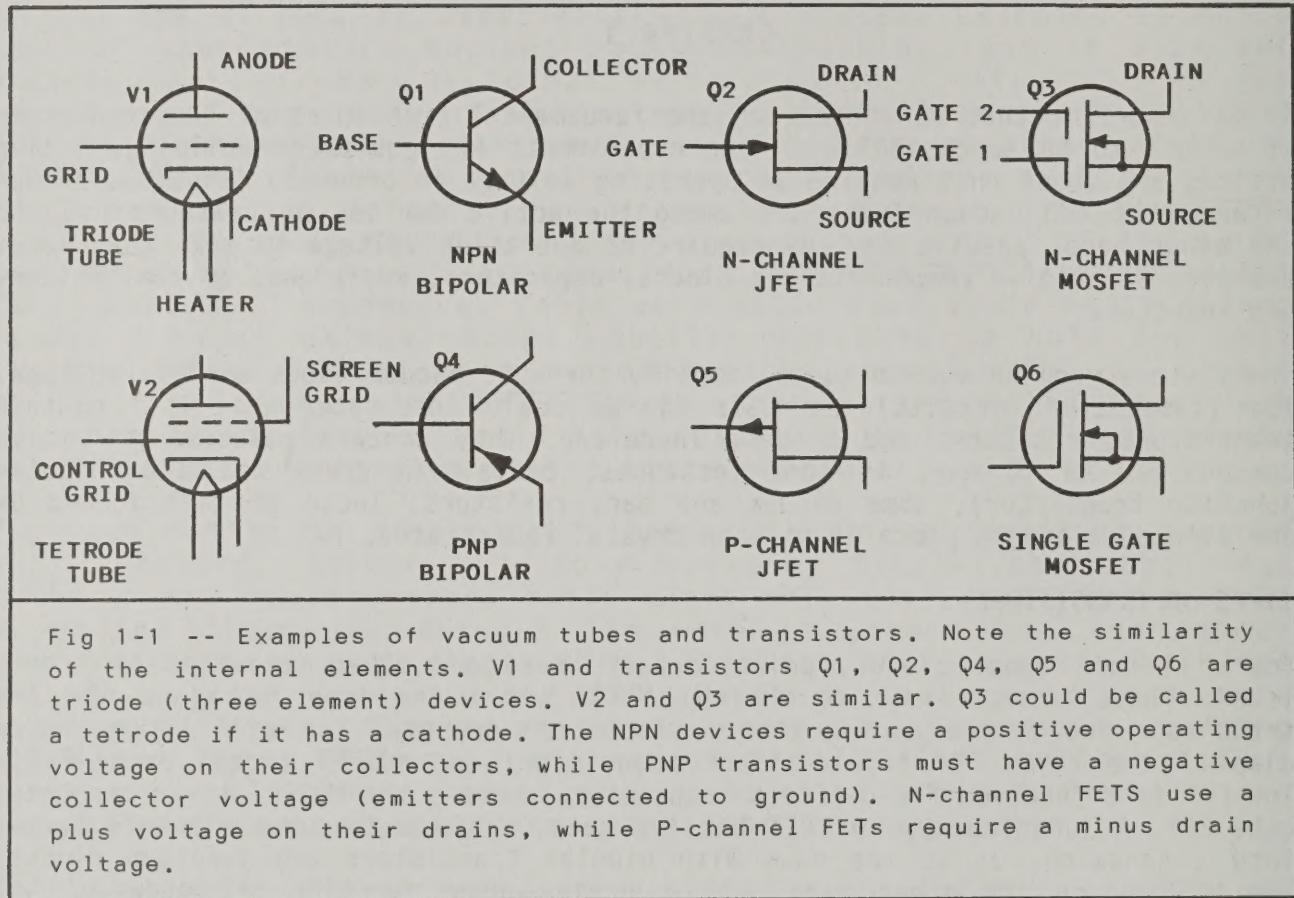


Fig 1-1 -- Examples of vacuum tubes and transistors. Note the similarity of the internal elements. V1 and transistors Q1, Q2, Q4, Q5 and Q6 are triode (three element) devices. V2 and Q3 are similar. Q3 could be called a tetrode if it has a cathode. The NPN devices require a positive operating voltage on their collectors, while PNP transistors must have a negative collector voltage (emitters connected to ground). N-channel FETs use a plus voltage on their drains, while P-channel FETs require a minus drain voltage.

NPN versus PNP

When designing a circuit from scratch we can use NPN or PNP bipolar transistors, irrespective of the available power-supply polarity. The same is true when we work with N-channel and P-channel FETs. For example, if we wish to use a negative power supply with a circuit that contains NPN transistors, we need only to keep the emitter circuitry above circuit ground, then feed the minus supply voltage to the emitter through the emitter components. Under this arrangement we must return the collector to chassis ground (positive in this example) through its components. The base resistor or other component that is normally grounded when NPN devices operate from a positive power supply must be connected to the minus voltage that feeds the emitter. PNP transistors can be used in the same manner by feeding the +V to the emitter circuit and returning the collector to ground. You may treat N-channel and P-channel FETs in the same manner. Using this technique we can actually use NPN and PNP devices in the same circuit when only one power-supply polarity is available. Fig 1-2 illustrates how this may be done for NPN and PNP transistors. Knowledge of this practical procedure enables us to take advantage of bargain-price transistors. It also permits us to use whatever may be on hand in the workshop storage bins.

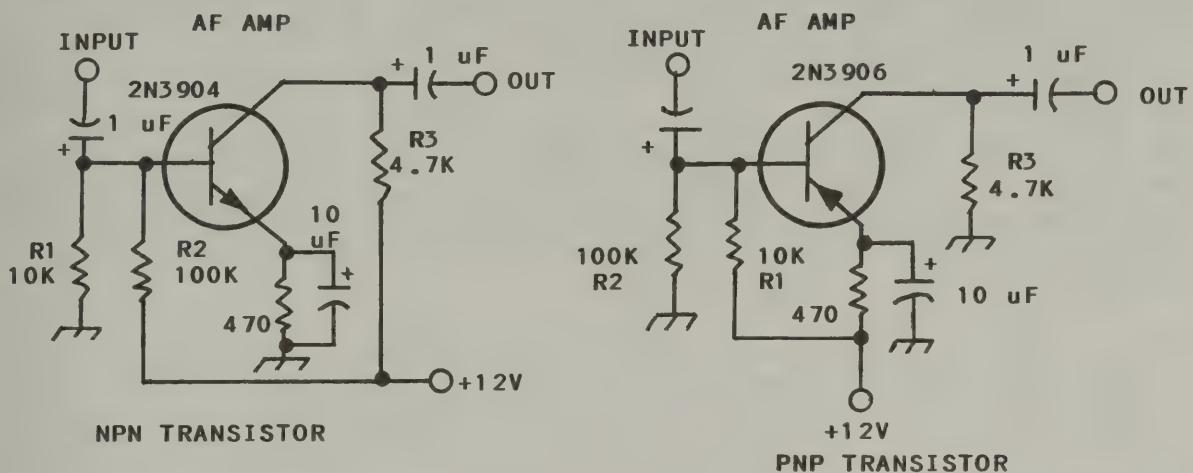


Fig 1-2 -- Illustration of how one may use either NPN or PNP transistors with a positive dc power supply. Note the relationship of R1, R2 and R3 to the two circuits. This same arrangement may be used when employing a negative power supply with NPN transistors. In this case the negative voltage is applied to the emitter of the NPN transistor and R3 is grounded.

TRANSISTOR RATINGS AND TERMS

Many newcomers to electronics experimenting are baffled by the terms that relate to transistor performance. There are many terms to consider, but most of the symbols are not of special interest to amateurs. Don't let these terms scare you. For example, we often hear the term "beta." This relates to the current gain of the transistor -- an important parameter when we design a circuit. There are two expressions for beta. One relates to the dc gain of the transistor (hFE). If we apply a small dc voltage to the base of the transistor, causing, say, 1 mA of base-emitter current to flow, we can now measure the collector current and learn what the dc gain (beta) is. If with 1 mA of base current we cause 100 mA of collector current to flow, we have a dc beta of 100. The higher the number the greater the transistor gain. There is also an ac (small signal) beta characteristic (hfe). This is generally the parameter that amateurs are concerned with. Let's use the 2N3904 in Fig 1-2 for an example. Motorola specifications list the hFE as 100 to 300 when there is 1 volt of VCE (collector-to-emitter) and an IC (collector current) of 10 mA. The greater the collector current the lower the hFE . Now, let's look at the hfe rating for a 2N3904. The manufacturer lists a beta spread of 100 to 400 when there is a 1-kHz signal applied to the base, and when the collector current is 1 mA with a VCE of 10 volts. The manufacturer is unable to state a specific hFE or hfe , hence the "beta spread" listing. This is because production techniques do not permit precise predetermination of the transistor characteristics for a given production run. Therefore, some transistors with the same numbers have slightly different performance traits.

A rather important term for transistors is "fT." This relates to the upper frequency limit of the device. This characteristic is known also as the "gain-bandwidth product." The 2N3904, for example, has an fT of 300 mHz. This means that the hfe is unity, or 1 at 300 MHz. The gain increases markedly as the operating frequency in MHz is lowered. In essence, the 2N3904 has no gain at

300 MHz. Most transistors will, however, perform as oscillators at or somewhat above their fT ratings. Efficiency is usually quite poor at and above the fT.

Maximum Ratings: We need to be ever mindful of the maximum safe ratings for all semiconductors. These are listed on the manufacturer's data sheets and in various transistor manuals. Let's once again use the 2N3904 as an example. The maximum VCEO (collector-to-emitter voltage with the base open) is +40. The maximum IC (steady collector current) is 200 mA. Maximum PD (total device power dissipation at 25°C [72°F]) is 1.5 watts. Now, let's learn what these ratings mean to us.

Transistors are likely to self-destruct if the foregoing ratings are exceeded, even for short intervals. We should never operate a transistor at its maximum ratings if we want it to endure. Excessive VCE can puncture the transistor junction (cause an internal short circuit) and too much IC can melt the junction, causing an internal open circuit. I recommend operate your transistors at half or less the maximum ratings. This allows ample margin for safety.

Keep in mind that the transistor VCE rises to twice the supply voltage when ac or RF energy is being amplified, owing to the sine-wave function. In other words, if we used a +24-V VCE for a 2N3904 audio amplifier, the VCE can swing as high as 48 volts at the peak of the sine wave! This exceeds the safe VCE by 8 volts. In certain other circuits (an amplitude-modulated Class C 2N3904, for example) the VCE can theoretically increase to four times the VCE!

The effective VCE can soar to very high levels when an amplifier stage breaks into self-oscillation or when an RF amplifier looks into a high SWR while it is operating at maximum power. Therefore, we must consider many aspects of the VCE when we select a transistor for a given application.

No transistor should be allowed to operate when it is hot to the touch. Heat is the no. 1 enemy of semiconductors. Reduce the operating parameters or use a heat sink when a transistor is more than comfortably warm to the touch. Likewise when working with diodes and ICs. Excessive heat usually indicates a design or application problem.

TRANSISTOR SATURATION

The expression "saturation" is a common one. What does it mean? Let us suppose that we have an RF amplifier stage in a solid-state transmitter. We need to obtain more output from this stage than we are getting. The typical approach by experimenters is to increase the driving power to the stage in question. We raise the driving power from, say, 100 mW to 200 mW, but there is no increase in output power from the inadequate stage. How can this be? The simple answer is that the transistor was already saturated at 100 mW of drive. Saturation means that no more output can be obtained when increasing the excitation. Vacuum tubes exhibit the same characteristic. The only way to increase the power output is to use a higher VCE, but at the risk of exceeding the safe ratings of the transistor. The lower the VCE the quicker saturation occurs. You may observe this phenomenon by building a small RF amplifier and measuring the output power with a scope or RF voltmeter as you apply and increase the excitation. You will find a point at which no further power output occurs with an increase

in driving power. This is the saturation point. Further increases in drive only cause increased IC and degradation of the output waveform. I suggest that you operate your amplifier stages just below the point where saturation occurs.

The term saturation is frequently heard with relation to bipolar dc switches. A bipolar switch is a transistor (NPN or PNP) that is used to switch dc or ac circuits in place of a mechanical type of switch. Diodes may be used as switches also. In the case of a bipolar transistor switch we need to apply forward bias (base voltage) in order to make the transistor conduct. When it is biased into full conduction (turned on) it can cause a relay to close or allow a supply voltage to pass through it to the desired circuit. Sufficient forward bias is applied during this process to cause the collector-emitter junction to function as a slightly resistive switch. In essence, the transistor is in saturation when it is completely turned on. If the bipolar switch is used to actuate a relay, it is referred to as a "relay driver." The relay field coil is placed between the transistor collector and the supply-voltage source. When sufficient collector current flows through the relay coil, the relay closes. It is not possible to obtain a finite resistance between the collector and emitter when a transistor switch is turned on. Generally, the effective junction resistance is a few ohms or less. Fig 1-3 shows three examples of dc switches.

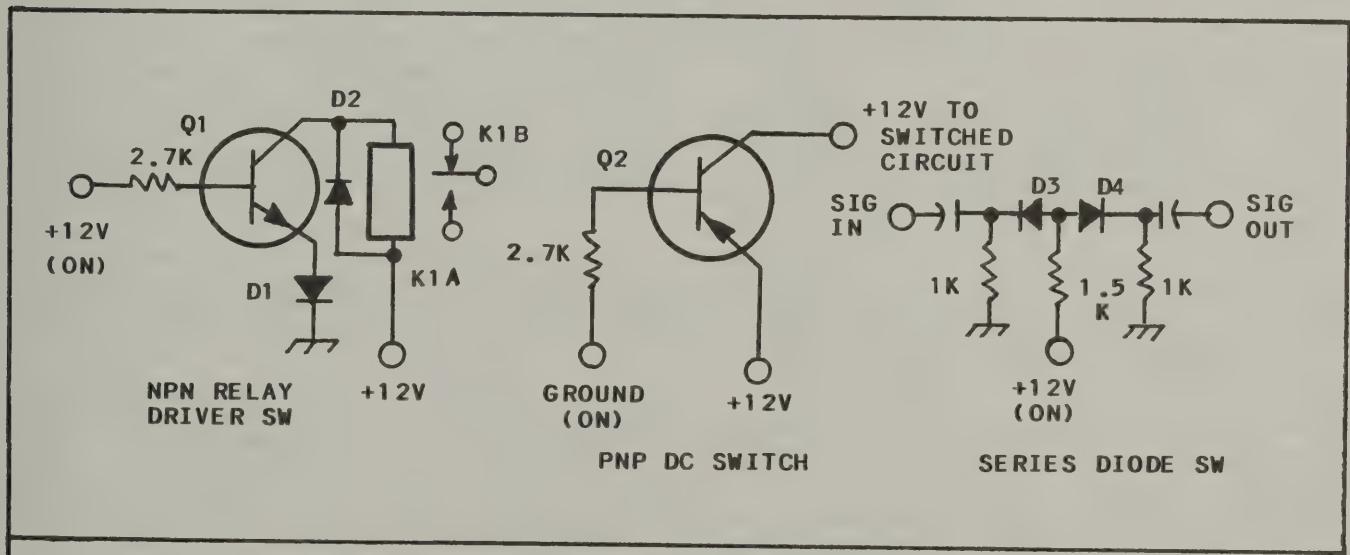


Fig 1-3 -- Examples of dc switches using bipolar transistors and silicon diodes. Q1 is a relay driver (NPN). Application of +12 V to the base causes the transistor to conduct, draw collector current and actuate 12-V relay K1. D1 provides bias that ensures fast turnoff of K1. D2 clips transients caused by the collapse of the K1 field coil. Q2 is a PNP switch that is turned on when the base is grounded. D3 and D4 provide a series dc switch that may be used in an audio or RF signal line. Application of +12 volts causes the diodes to conduct. 1N914 diodes are used in these circuits.

There are numerous variations of the above electronic switches. In all instances the semiconductor device must be turned on if it is to function as a switch. It can be considered as saturated when it is actuated.

TRANSISTOR POWER AMPLIFIERS

Bipolar and FET devices are commonly used as large-signal amplifiers from audio frequencies through the microwave spectrum. Class C operation is generally chosen for CW and FM amplification (nonlinear amplifier). Class A or AB is necessary for amplifying AM and SSB signals, owing to the need for linearity during the amplification process. A linear amplifier is one that faithfully reproduces the input waveform at the output of the amplifier. A linear amplifier that is functioning properly creates only minor distortion products (typically 30 dB or greater below the peak power output of the amplifier).

Bipolar transistors are available for use into the microwave region. Power FETs are at this time suitable as amplifiers into the VHF spectrum. The efficiency of both devices is on the order of 50-60 percent in Class AB service, but Class C efficiencies of 70 percent or greater can be achieved when these transistors are used as RF amplifiers. Tuned (narrow-band) power amplifiers are more efficient than broadband amplifiers. The increased bandwidth of the latter amplifier results in a trade-off in amplifier gain. Amplifier efficiency is determined by the ratio of the power dissipation of the stage versus the signal output power. For example, if we had a RF amplifier that produced 10 watts of output power, while having a VCE of 12 volts and a collector or drain current of 1.67 A (20 W), the efficiency of the stage would be 50 percent.

Amplifier gain is determined by the signal input power versus the signal output power. Suppose that our amplifier delivers 10 watts of output power. The driving (input) power needed to make this happen is 1.5 watts. The stage gain is 8.24 dB. This is obtained from $\text{Gain(dB)} = 10 \log P_1/P_2$, where P_1 is the higher power and P_2 is the lower power. Transistor power amplifiers can produce gains as great as 15 dB, which is dependent upon the device we select for a given circuit. High-gain power transistors require careful circuit design in order to prevent unwanted self-oscillation (instability).

Instability can be ensured by adopting a number of design measures, such as keeping all signal leads short and direct, using a PC board that has a ground plane (copper surface) on the component side of the board, and by using feedback.

Feedback reduces the amplifier gain by routing some of the amplifier output power back to the input circuit. This is known as negative feedback. Positive feedback, on the other hand, encourages self-oscillation or regeneration. The negative feedback is 180 degrees out of phase with the input signal. Positive feedback occurs when the feedback energy is of the same phase as the input signal. The greater the negative feedback used the lower the amplifier gain. Therefore, we must use only enough feedback to assure amplifier stability.

FETs versus Bipolar Transistors: Bipolar power transistors can deliver nearly their rated output power when the VCE is reduced somewhat below the rated value. By way of example, a 12-V transistor may be rated at 25 watts of output power. This same transistor might deliver 20 watts of output power at 9 or 10 volts. A power FET, on the other hand, loses output power much faster when the VDS is reduced by a few volts. Best results can be expected when a power FET is operated at its recommended VDS (drain-source voltage). A 24-V power FET may yield 15 watts of output power at 24 V, but at 12 V the output power can drop to only 5 or 6 watts with the same driving power. Greater output power may be obtained at 12 V by increasing the resting drain current (biasing), but the efficiency of the stage will be very poor.

Power FETs are more subject to VHF self-oscillation than are bipolar transistors. It is frequently necessary to use a "Q killer" or "de-Qing" device directly at the gate of the power FET. A miniature ferrite bead (850 μ u) may be slipped over the gate lead of the transistor. Alternatively, we can place a 10-ohm, 1/4-W carbon composition resistor in series with the input signal, directly at the transistor gate. In broadband amplifier circuits we can shunt a capacitor from the transistor drain to ground. It should have an XC (capacitive reactance) of a least four times the drain impedance. Examples of the above methods for suppressing VHF oscillations are shown in Fig 1-4.

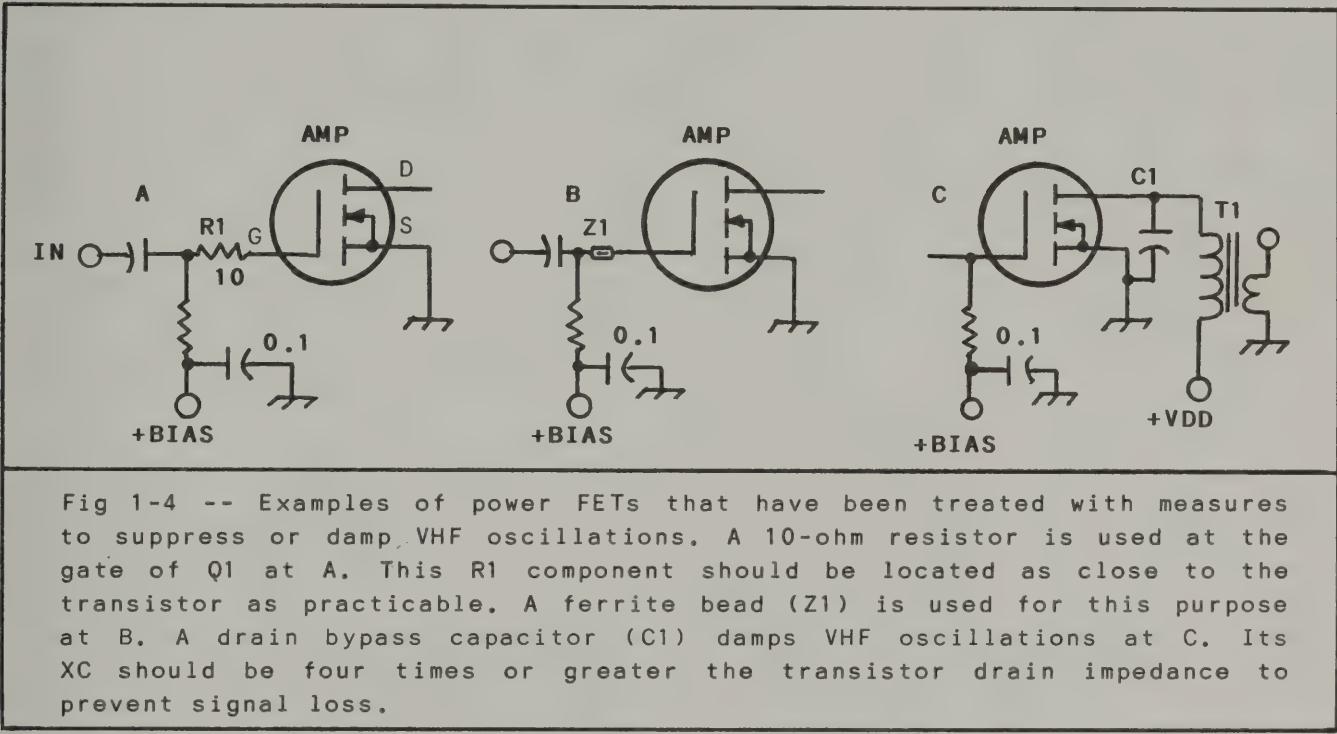


Fig 1-4 -- Examples of power FETs that have been treated with measures to suppress or damp VHF oscillations. A 10-ohm resistor is used at the gate of Q1 at A. This R1 component should be located as close to the transistor as practicable. A ferrite bead (Z1) is used for this purpose at B. A drain bypass capacitor (C1) damps VHF oscillations at C. Its XC should be four times or greater the transistor drain impedance to prevent signal loss.

R1 of Fig 1-4A may be used in combination with C1 of Fig 1-4C. In a like manner you may use Z1 and C1 of Fig 1-4 to VHF-suppress an amplifier stage. These techniques introduce negligible signal loss below 30 MHz. If the drain impedance of the FET in Fig 1-4C is, say, 15 ohms, and the operating frequency is 7 MHz, C1 will be 379 pF or less. The approximate drain impedance is obtained from $Z(\text{ohms}) = V_{DS}^2/2P_0$, where P_0 is the amplifier output power in watts. C1 of Fig 1-4C can then be calculated from $C(\mu\text{F}) = 2\pi \times 4ZD \times f(\text{MHz})$, where π is 3.14 and ZD is the calculated drain impedance. If C1 is used in a multiband amplifier, choose the C1 value for the highest operating frequency.

FETs have a high gate impedance (1 megohm or greater), whereas bipolar transistors exhibit a low input impedance -- usually less than 10 ohms in an RF power amplifier. The FET gate impedance is set by the input transformer or the gate resistor in the bias line. Typically, the gate impedance is made low intentionally by the external components at the amplifier input. Forcing the impedance to a low value aids amplifier stability but requires greater driving power than if the driver stage were looking into a relatively high impedance. A 50- to 100-ohm forced impedance is typical. We need sufficient driving power to cause a signal-voltage swing of roughly 30 volts peak-to-peak at the gate. This voltage ensures sufficient excitation to make the FET draw drain current and amplify. The gate of a power FET draws only microamperes of current, which makes it a simple task to devise a resistive bias network.

Whereas bipolar transistors have a beta (h_{FE}) rating to indicate their gain capability, FET gain is expressed in siemens or mhos. These terms relate to the transconductance of the FET. Vacuum-tube transconductance is expressed also in mhos. This is because a FET or tube is a voltage amplifier, whereas the bipolar transistor amplifies current. A mho or siemens is a unit of conductance (or admittance) in the international system of units (SI). The greater the FET transconductance the greater its gain potential.

Internal Capacitance -- FETs vs BJTs: Bipolar transistors that are used in RF power amplifiers exhibit a substantial amount of input capacitance. This capacitance varies with the operating frequency, and becomes greater as the operating frequency (in MHz) is lowered. A given transistor may have as much as 3000 pF of input capacitance at 1.8 MHz. This capacitance, along with the smaller output capacitance, varies with the operating voltage, current and frequency. The capacitance change makes it difficult, but not impossible, to design a good broadband amplifier that uses a feedback network. This varying capacitance causes a nonlinearity of transistor action that helps to generate harmonic currents (varactor action). It is not unusual to find the second and third harmonic energy only 10 or 15 dB below the peak fundamental frequency at the transistor collector. A correctly designed low-pass filter at the amplifier output can resolve this problem by attenuating the harmonic currents by 40 dB or greater. The larger the transistor internal geometry the greater the input capacitance, as a general rule.

Power FETs do not undergo these large changes in capacitance. RF power FETs have a relatively low input and output capacitance compared to BJTs. The internal capacitance remains relatively constant, as is the case with vacuum tubes. This feature minimizes harmonic generation and makes it a simple matter to design an effective feedback network for a broadband amplifier. Furthermore, the FET gain remains the same up to its rated upper frequency. On the other hand, the theoretical gain increase for BJTs is approximately 3 dB for each octave lower in frequency. Thus, if the BJT has a gain of 10 dB at 14 MHz, the gain becomes 13 dB at 7 MHz, and so on. Instability becomes a major consideration when using VHF BJTs in the HF spectrum, owing to the increased gain.

The FET input capacitance is expressed as C_{iss} . Output capacitance is identified as C_{oss} . The symbol for transconductance is g_{fs} . The expression for input capacitance of a BJT is C_{ib} , and the output capacitance symbol is C_{ob} . These values are generally stated on the manufacturer's device data sheet.

High values of input capacitance make it difficult for us to design input-matching networks for the upper part of the HF spectrum. The capacitance needs to be absorbed in the matching network, which often results in unworkable values of coil inductance (a fraction of a microhenry in a worst-case example).

DEVICE SELF-DESTRUCTION CAUSES

The most common cause of BJT failure (assuming normal operating voltage and current) is thermal runaway. This is a function of heat, and usually occurs if a power transistor does not have adequate heat-sink area. As the BJT operating temperature (internal) increases, so does its gain. This increases the collector current, which generates more heat and gain, thereby causing the ultimate destruction of the transistor. The net result is an open or melted

transistor junction. Excessive collector-base or collector-emitter voltage can also destroy a BJT. Excessive voltage punctures the device junction and causes an internal short circuit. An ohmmeter can be used to determine whether excessive voltage or current destroyed a transistor.

There is also a phenomenon called "beta degradation." In this case the BJT gain falls off over a period of time. The usual cause is excessive base current that is brought on by too much excitation. Too great a level of base-emitter voltage, which accompanies excessive drive, may also result in beta degradation.

Power FETs are somewhat more fragile than BJTs with respect to excessive voltage or gate current. Transients that appear on the input signal can puncture the gate insulation instantly. In a like manner, voltage spikes or transients on the FET drain can cause instant destruction. Most modern power FETs that are used in switching applications contain a built-in drain-source Zener diode to protect the transistor from voltage peaks. Some power FETs have a Zener diode from gate to source as well. A designer may add this form of voltage clamping external to the FET, if these diodes are not present inside the transistor. Protective Zener diodes do, however, add to the transistor input and output capacitance.

A power FET is virtually immune to thermal runaway -- a plus feature. They are unlikely to self-destruct in the presence of high SWR, which is not true of all BJTs. Some BJTs have internal SWR protection which allows them to survive when the load is mismatched from an open to a shorted condition. They are called "ballasted" transistors. The substrate of the transistor contains many BJTs in parallel, but their emitters are returned to a common emitter pin through tiny 1-ohm resistors within the transistor. These emitter resistors equalize the currents of the various transistors so that no one BJT "hogs" the current. Current hogging causes what is known as "hot spotting," which can destroy the current-hogging transistor or transistors on the substrate.

When you work with any power transistor it is important that you provide ample heat-sink area. It is best that no transistor become too hot to touch with your finger. If the device should become extremely hot to the touch, increase the area of the heat sink. A cooling fan may be used to lower the temperature of the transistor and its heat sink, should your heat sink be too small for the application.

Another potential threat to a power transistor is improper mounting of the device on a heat sink or PC board. The microstrip base, collector and emitter leads on some power transistors should remain at a right angle to the transistor body when it is mounted. In other words, don't bend these leads up or down when soldering them to the PC board. This causes stress on the body (header) of the transistor, which can damage the internal workings of the device when it is hot (operating). Do not over-tighten the nut on the transistor mounting stud. It should be snug against the heat sink, but never drawn down severely. A thin layer of silicone grease (silicone and zinc oxide) should be spread over the area of the heat sink where the transistor mates with the heat sink. This helps to ensure the proper transfer of heat from the transistor to the heat sink.

Small transistors, such as those in TO-92 or TO-18 cases, can be cooled by using small press-on heat sinks, or tabs of copper or aluminum that are affixed

to the transistor body by means of epoxy glue. Clamp the metal tab to the transistor body before applying the cement. This ensures a firm bond between the two mating surfaces.

DIODES

We find ourselves working with a wide variety of diodes when building amateur projects. The most common ones are those that function as rectifiers in power supplys. Small-signal diodes, such as the 1N34A and 1N914 types, find regular use as signal detectors, mixers and frequency multipliers. We also use a number of Zener diodes to provide various regulated dc voltages. VVC (voltage-variable capacitance) diodes or varactors are useful in place of mechanical tuning capacitors in a number of amateur circuits. Varactor diodes are used also as frequency multipliers that provide substantial output power. Fig 1-5 shows the symbols for various common diodes.

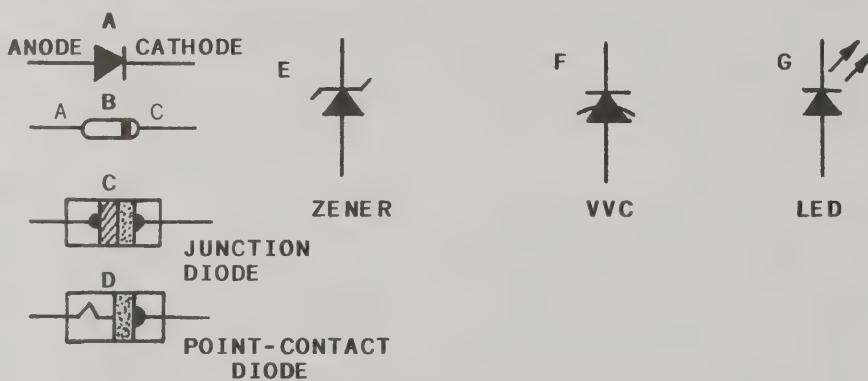


Fig 1-5 -- Symbols for various common semiconductor diodes. The example at A shows the symbol for junction and point-contact diodes. Drawing B shows a diode in its case with the band that indicates the cathode end. A Zener diode is seen at B, a VVC diode at F and a light-emitting diode (LED) at G.

You can see that the junction diode (Fig 1-5C) consists of a sandwich of two semiconductor materials (P and N) with the electrodes sintered to the wafers. This represents the inner structure of silicon diodes. Fig 1-5D illustrates the inner structure of a germanium point-contact diode. It is similar to a galena crystal and a catswhisker from the days of crystal-detector radios. The barrier voltage (voltage required to make a diode conduct) for silicon diodes is approximately 0.7, whereas the germanium diode conducts at roughly 0.3 to 0.4 volt. The germanium diode, such as the 1N34A and 1N60A, are more sensitive than are silicon diodes (such as 1N914) in circuits that have low signal voltage. The silicon small-signal diode can accommodate higher current and voltage than can a small germanium diode.

Diode Ratings: Rectifier diodes are rated for maximum PRV (peak reverse voltage) or PIV (peak inverse voltage) and current. Excessive PRV can ruin a diode by causing a short circuit through the diode junction. Too much current (and heat) will melt the junction and cause an open circuit.

Zener diodes are rated for the regulated voltage they provide, such as 9.1 V. They also carry a wattage rating, such as 400 mW, 1-W, etc. The wattage

rating indicated how much current versus voltage the device can tolerate without breaking down. The higher the wattage rating the greater the maximum current of the circuit to be regulated.

VVC diodes are rated for maximum voltage and the capacitance change they undergo as the operating voltage is varied. Maximum diode capacitance occurs at minimum reverse voltage (a positive voltage is applied to the diode cathode, and the anode is grounded). Various ranges of capacitance may be had by selecting the appropriate VVC diode. These diodes are rated also for Q (quality factor) versus the operating frequency. The diode Q should be substantially higher than the loaded Q of the tuned circuit with which it is used.

LEDs have a barrier voltage of 1.5 and a maximum current of roughly 20 mA, depending upon the type of LED used. The smaller LEDs usually operate with less than 10 mA of current. LEDs are available in red, green, yellow and clear lenses. They are useful also as voltage-reference devices if we take advantage of their 1.5-V barrier characteristic. You may think of them in this application as low-power 1.5-V Zener diodes, but with the operating voltage applied to the anode rather than the cathode (positive voltage is applied to the cathode of a Zener diode).

Hot Carrier Diode: This is a special type of diode that has a low barrier voltage (0.25) and is ideally suited to high-speed switching and mixer applications. This semiconductor is known also as the Schottky diode. The forward and back resistances of hot-carrier diodes from a given production run are nearly identical when we compare the diodes with an ohmmeter. This makes them ideal for use in balanced mixers and modulators, where matched diodes ensure the best circuit balance. Also, they conduct at much lower signal levels than is true when using the 1N914 type of silicon diode.

PIN Diodes: These diodes are used as switches in low-level signal lines from VHF through the microwave spectrum. They act as variable resistances, which enables them to switch signals on and off (such as when they cause a short circuit to occur across a waveguide). Their characteristics do not make them especially applicable to HF operation.

Gunn and IMPATT Diodes: Gunn diodes are used as low-power oscillators from, say, 5 to 100 GHz. They can provide output powers of approximately 100 mW.

IMPATT diodes function in a manner similar to that of a klystron. IMPATT diodes may be used as microwave oscillators to produce up to a few watts of output power.

Photodiode: These diodes have a junction that functions as a photo detector. When exposed to light they cause a current to flow through a circuit that is external to the diode. This current may be used to actuate a switching transistor that can turn on a relay or actuate another circuit. These diodes have a transparent case or lens that permits light to strike the diode junction. Photodiodes are designed for high-speed switching, low noise and low leakage current.

Photovoltaic Diode: This diode is known as a "solar cell." This diode is also light-sensitive (similar to a photodiode), and a single solar cell can produce a 0.5-V output. The larger the disc- or rectangular-shaped cell the greater the output current. These large silicon wafers are used in series in solar-

electric panels. Typically, 36 cells are used to develop 18 volts of panel output at peak sunlight. This voltage may then be passed through a "floating regulator" to provide 12-14 V of dc output. The larger solar cells can deliver up to 1.5 A of output current. Solar panels may be used in series to obtain higher voltages. In a like manner we can place them in parallel to obtain higher output current.

GaAsFETs

One of the more modern transistors is the GaAsFET. This field-effect transistor is intended for RF amplification at UHF and microwaves. A specific GaAsFET may, for example, deliver 11 dB of gain at 5 GHz with a 1.3-dB noise figure. This transistor is made with gallium arsenide (used also in LEDs and some microwave diodes) rather than the more common germanium or silicon. The FET gate is made from gold or aluminum. GaAsFETs are available for use up through the Ku-band (15.3 GHz). They are ideal for amateur low-noise, high-gain preamplifiers or converter RF amplifiers from VHF upward. Both low- and high-power GaAsFETs are available. The large-signal devices have nearly as low a noise figure as the smaller units and they have a much greater dynamic range (strong-signal handling capability versus overload and IMD degradation).

INTEGRATED CIRCUITS

You will observe that two types of integrated circuits (ICs) are available to you. Linear ICs are those that can amplify signal (ac or RF) energy. An example of a linear IC is the popular 741 op amp or the MC1350P IF/RF amplifier. LSI (large-scale integration) ICs have much bigger packages than the common 8, 14 and 16 pin DIP (dual in-line package) ICs, and they have many more pins than typical ICs. Linear ICs perform many jobs other than simple signal amplification: they are used as mixers, balanced modulators, dc amplifiers, comparators, and so on.

The other IC type is a digital or logic IC. These ICs are used in frequency counters, computers, speech synthesizers and a host of similar devices that require logic circuitry.

Subsystem ICs: A number of complex ICs are available to us for building compact equipment. A subsystem IC is one that contains most of the semiconductors, along with internal resistors and capacitors, to represent the heart of a given circuit. There are AM and FM receiver subsystem ICs, plus some that can be used as the heart of an SSB receiver. Similarly, we can purchase an FM transmitter subsystem IC, or one around which to build a frequency counter. These advanced ICs require some external components, but permit us to construct rather tiny assemblies of the "Dick Tracy wrist-radio" variety.

IC Hardware: An IC may be "hard wired" (soldered directly to a PC board) into a module, or we can use IC sockets for mounting the chips (common name for an IC). Low-profile IC sockets are almost mandatory when working with linear ICs. A low-profile socket is one that has short pins and a thin plastic frame. These sockets minimize the potential for instability and other maladies that result from excessive lead length in circuits that have high gain. The larger wire-wrap types of IC sockets are generally okay for use in digital circuits.

Heat sinks are available for ICs. These gadgets are affixed to the top of the

IC by means of epoxy cement. They help to minimize excessive heat when used with ICs that dissipate substantial power. Audio-output ICs that deliver several watts of power are typical of integrated circuits that need to be cooled to prevent damage or malfunction.

MMICs: Monolithic microwave ICs represent a significant advance in IC design. These smaller than typical chips are low-noise amplifiers that operate in the VHF, UHF and microwave spectrum. They are in small packages and have flat leads that exhibit low inductance, which is essential to proper performance at the higher frequencies. They are broadband amplifiers capable of gains as great as 25 dB, and contain internal resistors that minimize the component count outside the IC. MMICs are designed to have a characteristic input and output impedance of 50 ohms. This makes them especially adaptable to PC boards that use 50-ohm strip-line copper elements. MMICs may be cascaded (used in series) to develop microwave amplifiers that exhibit 40 dB or greater gain, but this requires special care in laying out the amplifier PC board to prevent unwanted self-oscillation. Effective power-lead decoupling is also mandatory for each amplifier section in order to avoid instability.

CHAPTER SUMMARY

Some amateurs are afraid to experiment with circuits that use transistors and ICs, especially those hams who grew up in a vacuum-tube world and have had no formal education respective to semiconductors. "Learning by doing" is an acceptable trade-off for those of you who have not gone to school to learn solid-state device theory and application. If you accept transistors as current amplifiers rather than voltage amplifiers, you're on the road to being a skilled experimenter while using today's devices. Most of the theory you need can be found in **The ARRL Handbook** and other League books.

I suggest that you begin with some simple circuits, such as a speech amplifier for a microphone, a meter amplifier, an IC audio amplifier or a simple receiver of the DC (direct-conversion) variety. The more you work with solid-state devices the greater your knowledge and confidence will be. You will wonder why you waited so long to get started!

This chapter presents a simple overview of transistors, ICs and diodes. An entire book could be devoted to the topics in this chapter, but page space is not available to enlarge upon these topics. You can start on your adventures as an experimenter with nothing more than a small regulated +12-V dc power supply, a VOM and some hand tools. In other words, you need not have a laboratory that is equipped with expensive test equipment.

Circuit boards also tend to frighten some would-be experimenters. Again, this should not be a barrier for you. It is easy to make your own PC board (see later chapter), but you can avoid PC boards by using perforated phenolic board material, or "ugly construction" that calls for mounting the parts somewhat helter-skelter on a small piece of blank circuit-board material. Multilug solder terminals may then be used to support the various components. The copper foil on the board material serves as the common point for parts that are grounded on one end.

Glossary of Chapter Terms

Active device - A component, such as a transistor, that requires an operating voltage.

Ballasted transistor - A transistor that has built-in SWR protection, consisting of many bipolar transistors in parallel on a silicon substrate, and with each transistor emitter returned to a common pin through a 1-ohm internal resistor.

Beta - The measure of the approximate gain of a transistor. HFe is the dc current gain and hfe is the small-signal current gain.

Beta spread - The range of transistor beta for a given type of transistor. Each transistor from a production exhibits a slightly different beta, owing to nonuniformity of characteristics during the manufacturing process.

BJT - Bipolar junction transistor.

Broadband - A circuit or component characteristic that permits it to function over a wide spectrum of frequency without being tuned. A broadband transformer, for example, might provide uniform performance from 1.8 to 30 MHz.

C_{iss} - The characteristic input capacitance of a transistor.

C_{oss} - The characteristic output capacitance of a transistor.

Decoupling - A method for isolating an active stage from other circuits nearby. Decoupling helps prevent signal voltage from migrating to other parts of a circuit via the voltage supply line. Normally done with a network consisting of a resistor and capacitors, or an RF choke and capacitors in the supply line to the active devices.

Efficiency - A measure of merit based on the power dissipated by an amplifier stage versus the useful output power, expressed in percentage.

Electronic switch - Also known as a solid-state switch. A diode or transistor that is used in place of a mechanical switch in a dc or RF line.

Feedback - An amplifier output voltage that is fed back to its input circuit intentionally or by accident. Used intentionally in oscillator and broadband amplifier solid-state circuits.

FET - Field-effect transistor.

Forward resistance - The internal resistance of a diode measured from the anode to the cathode. Back resistance is measured from cathode to anode.

f_T - The upper useful frequency limit of a transistor. The point at which the transistor gain is unity or 1. Associated with grounded-emitter use. Alpha represents the upper frequency limit for a grounded base transistor.

GaAsFET - A field-effect transistor that uses gallium arsenide for the inner elements. Noted for its high gain and low noise figure at VHF, UHF and microwaves.

Gain - A measure of the increase in signal power or voltage when amplified by a transistor, IC or vacuum tube. Frequently expressed in dB or dBm.

gm - Symbol for transconductance. Presently symbolized by SI (siemens). Also called mu. Previously expressed in mhos or micromhos.

Gunn diode - A special solid-state diode used at UHF and microwaves as a low-power oscillator.

Heat sink - A metal heat-conducting device that is attached or bonded to a power diode, transistor or IC to help lower the internal temperature of the semiconductor during operation.

hfe - Symbol for the ac current gain of a BJT.

HFe - Symbol for the dc current gain of a BJT.

Hot carrier diode - A VHF, UHF, microwave diode that has a fast switching time and a low barrier voltage (0.25 V).

IMPATT diode - A microwave diode used as an oscillator to produce up to a few watts of output power.

Instability - A condition that occurs in an amplifier stage, causing it to self-oscillate on one or more random frequencies. Caused by unwanted feedback.

JFET - Junction field-effect transistor.

Linear IC - An integrated circuit that is used for amplification of dc and ac voltages, as a mixer, balanced modulator or oscillator.

Logic IC - Integrated circuit used in logic circuits, such as computers and frequency synthesizers.

MOSFET - Metal oxide field-effect transistor. Unlike a JFET, which does not have an insulated gate. The MOSFET has a thin layer of silicon oxide between the gate and the remainder of the inner components.

MMIC - Monolithic microwave integrated circuit. Characterized by high gain and a low noise figure at VHF, UHF and microwaves.

Narrowband - Opposite of broadband. A component or amplifier that amplifies a narrow band of frequencies and must be tuned for each operating frequency. A narrowband transformer, for example, is tuned to a specific frequency.

Parameter - An operating characteristic of a circuit or component.

Passive circuit - Opposite of an active circuit. It requires no operating voltage and causes a loss (insertion loss) in signal power. A harmonic filter or a diode mixer can be called a passive device.

PD - Power dissipation in watts or milliwatts of a transistor. Usually relates to maximum safe dissipation at 25° C (72° F).

Photodiode - A diode with a clear glass case or lens that is sensitive to light and produces an output current when exposed to light.

Photovoltaic cell - A thin wafer of silicon that is a light-sensitive diode which produces 0.5-V output in bright sunlight. Known also as a solar cell.

PIN diode - A special diode used for signal switching at VHF, UHF and micro-waves. Operates on the principle of variable internal resistance.

Power FET - A MOSFET designed for large-signal amplification in watts.

Q - Figure of merit for a component. The greater the Q the lower the circuit loss and the narrower the potential bandwidth, as with a high-Q tuned circuit or quartz crystal.

Q killer - Slang expression for a device (usually a resistor or ferrite bead) used to spoil the Q of a specified circuit or component to discourage circuit instability. Known also as a D-Qing device.

Saturation - A condition that occurs when increased excitation of an amplifier produces no corresponding increase in amplifier output.

Siemens (SI) - Current term for expressing gm or transconductance.

Solid-state switch - A semiconductor device used in place of a mechanical switch to turn a circuit on or off by means of a control voltage applied to the solid-state switch.

Thermal runaway - A self-destruct condition that can befall a BJT when it becomes excessively hot. Gain increases with heat, causing the collector current to rise and produce more heat until the transistor burns out.

V_{CC} - Dc supply voltage to a BJT circuit.

V_{CE} - Operating voltage measured from a BJT collector to emitter.

V_{DD} - Supply voltage to the drain of an FET stage.

V_{DS} - Voltage measured from a FET drain to its source.

Zener diode - A diode used as a voltage regulator. Positive dc voltage is applied to the cathode of the Zener diode through a series resistor. The anode is grounded.

TRANSISTOR APPLICATIONS

CHAPTER 2

This chapter offers basic "recipes" for a number of common circuits that you will be working with. Many of the circuits may be used with other circuits in this book, thereby permitting you to build a composite unit of your choice. All of the circuits in this section are practical ones. The components have assigned values that allow you to duplicate these circuits with ease. None of the values listed are especially critical. Departures of 20% from the designated component values will not impair the performance of these common circuits.

AUDIO AMPLIFIERS and PREAMPLIFIERS

A good starting point for your first-time adventures with semiconductors is to build some audio amplifiers that use BJTs and FETs. Familiarity with these simple circuits will aid you later on when you work with other circuits that use transistors. Choosing the right component values is the most important aspect of good performance. Biasing a transistor is perhaps the most critical of the design considerations. Too little forward bias (base voltage) results in low stage gain. Too much bias can saturate a transistor and render it useless. In fact, excessive base voltage can cause transistor failure by making it draw excessive current, which can destroy the transistor junction.

It is important also, when working with audio amplifiers, to select component values that provide the desired audio response for the amplifier. As a general rule, the greater the capacitance of the coupling and emitter bypass capacitors the lower the frequency response. The "lows" can be rolled off by using small values of capacitance in these parts of the circuit. The high-frequency response can be limited also. This is done by using correct values of bypass capacitor from either the transistor base or collector to ground. In effect, we have a crude form of tone control when we do this. The desired capacitor values can be determined experimentally by listening to the output of an audio amplifier with headphones while having someone speak into a microphone that is connected to the amplifier input circuit. A more precise procedure for this experiment is to use a variable-frequency audio generator and an oscilloscope. The generator is set at the desired upper frequency limit of the amplifier, and various values of bypass capacitors are shunted from the transistor base, collector or drain until the highs are rolled off to your liking. The same procedure may be used when you select the coupling-capacitor values, along with those for the emitter or source bypass capacitors. I will not burden you with the equations that design engineers use for shaping the audio response of an amplifier.

It is important that you select a collector or drain resistor that allows a fair amount of voltage drop. I like to have approximately half the dc supply voltage at the collector or drain of the audio amplifier. In other words, if the VCC is 12 volts, select a resistor that causes the VCE or VDS to be roughly 6 volts. This allows the signal voltage (audio sine wave) to swing beyond the

VCC value without the peaks flattening and causing distortion. Always allow for plenty of collector-voltage swing when the stage is amplifying. The collector or drain load resistor can be determined by observing the collector current, then applying Ohm's Law ($R = E/I$, where E is the desired voltage drop and I is the collector or drain current in amperes). Fig 2-1 shows a number of basic audio preamplifiers with practical component values. These may be used, as shown, with operating voltages from 9 to 15 without changes in the parts values.

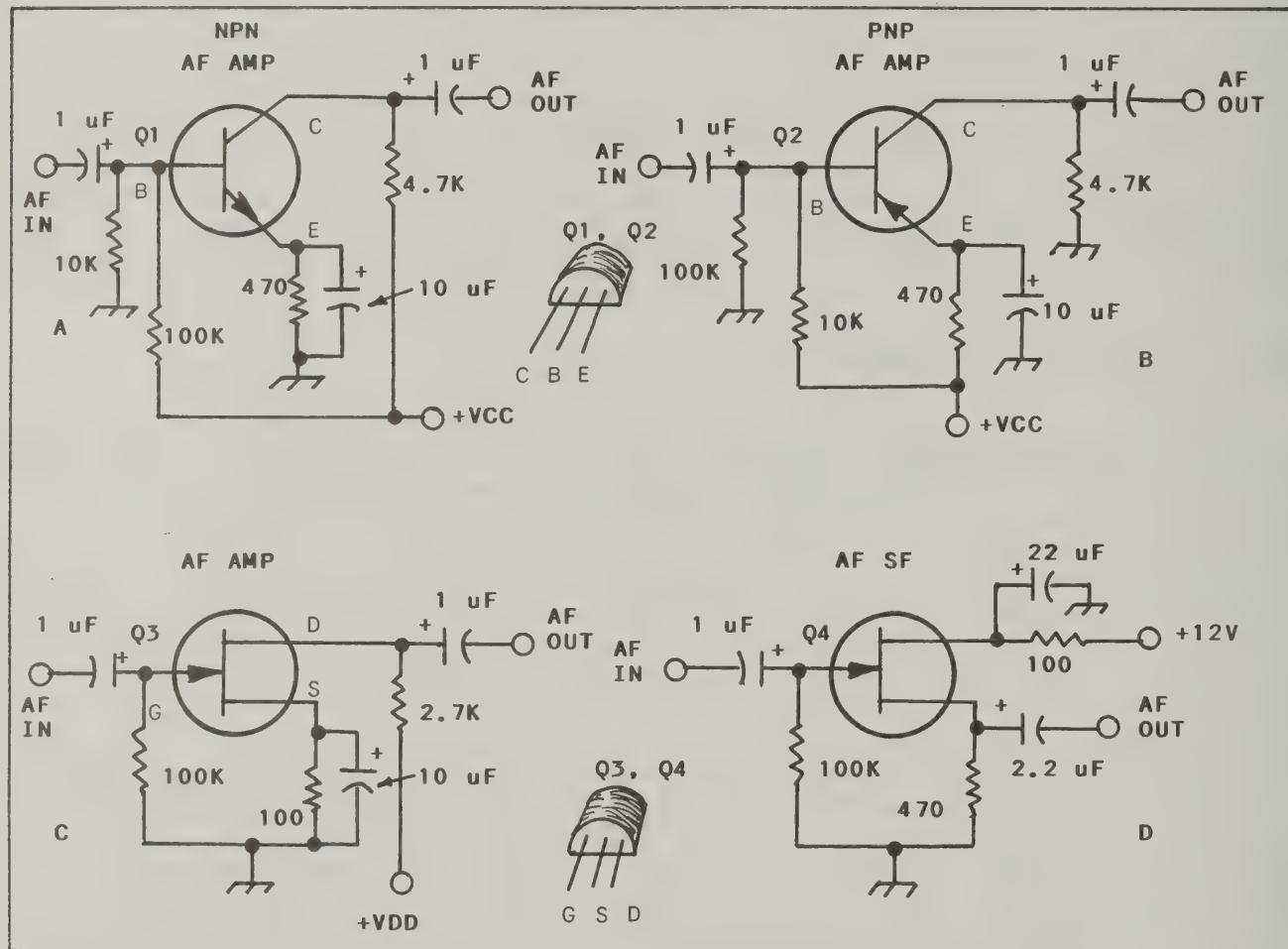


Fig 2-1 -- Examples of low-level audio amplifiers or preamplifiers. Q1 may be any common transistor such as 2N3904, 2N2222 or 2N4400. The input impedance of the amplifiers at A and B is on the order of 600-1000 ohms, making them suitable for use with 600-ohm mics. The output impedance is set by the collector resistor (4.7K ohms). The circuit at B shows how to use a PNP transistor (2N3906, 2N4403, etc.) with a positive supply voltage, as discussed on page 2. The stage gain at A and B is 10-15 dB typically. A JFET (MPF102 or equivalent) is used at C to provide a high-impedance input for hi-Z mics, such as the D-104 (50K ohms). The stage gain (C) is roughly 10 dB. A source follower is shown at D. The FET acts as an impedance transformer (hi- to low-Z) which permits the use of a hi-Z mic (such as D-104) with a low-Z (600 ohm) mic input circuit. The voltage gain (actually a slight loss) of the source follower is 0.9. The component values listed above are for voice communication and provide maximum response from 300 to 3000 Hz.

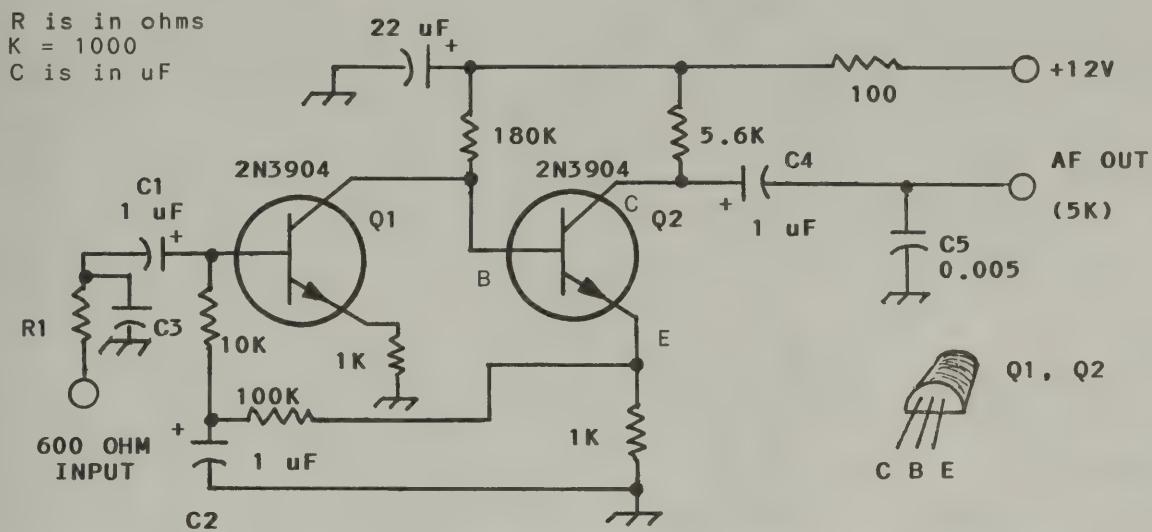


Fig 2-2 -- Example of a compound or direct-coupled audio preamplifier. The values for C1, C2 and C4 are chosen for the desired low-frequency rolloff. The greater the capacitance the lower the frequency response. The values listed above are for attenuation below approximately 500 Hz. C5 may be added to roll off the higher audio frequencies. Values between 0.001 and 0.1 μ F are suitable for speech-frequency limits of 5000 to 2000 Hz, respectively. R1 (1K) and C3 (0.005 μ F) may be added to prevent RF energy from entering the amplifier. A 1-mH RF choke may be used in place of R1 to preserve the 600-ohm input impedance.

Fig 2-2 shows a high-gain compound audio preamplifier that can be used as a mic booster, audio-input stage for a transmitter or a low-level audio amplifier in a receiver. It will perform well from +9 to +15 volts and provides a voltage gain (AV) of roughly 40 dB at +12 volts. You may increase the low-frequency response to 10 Hz by changing C1, C2 and C4 to 10 μ F. This low-noise preamp is temperature-stable because of the feedback circuit used. R1 and C3 may be used to filter unwanted RF energy from the mic lead. R1 will change the input impedance to 1600 ohms, but it will remain essentially 600 ohms if you substitute a 500- or 1000- μ H RF choke for R1. Transistors such as the 2N2222A and 2N4400 may be used in place of the 2N3904s shown in Fig 2-2. Any low-noise NPN transistor with an hfe equivalent to or greater than that of a 2N3904 is suitable for this circuit.

Output bypass capacitor C5 is used to shape the frequency response of the amplifier by rolling off the higher audio frequencies. This may be done experimentally while monitoring the audio output with headphones while speaking into the mic. An audio sine-wave generator and a scope can be used for this experiment to obtain more accurate shaping of the amplifier response. C5 also serves as an RF bypass capacitor (RF energy can sneak in the back door also). R1, C1 and C5 may be applied to most audio preamps when RF energy disrupts performance. This is evidenced by audio squealing and distortion.

Fig 2-3 shows a compound audio amplifier that is patterned after a circuit that was presented in the **RCA Transistor, Thyristor & Diode Manual**. It has a high input impedance (55K ohms), which make it ideal for high-Z mics such as the

D-104 and other units with crystal or ceramic elements. The amplifier output is 1 volt RMS (root mean square) across a 250-ohm load impedance. The gain is selectable. Gains as great as 166 are possible with this circuit (see chart in Fig 2-3). As the gain increases the input impedance decreases. This amplifier is temperature and voltage stable like the one in Fig 2-2, owing to the feedback circuit from the emitter of Q2 to the base of Q1. This circuit is excellent for use as an impedance transformer between an high-Z mic and a low-Z audio amplifier. Most modern SSB and FM transceivers have a low-Z mic input circuit. The Fig 2-3 preamp permits us to interface a high-Z mic with these rigs.

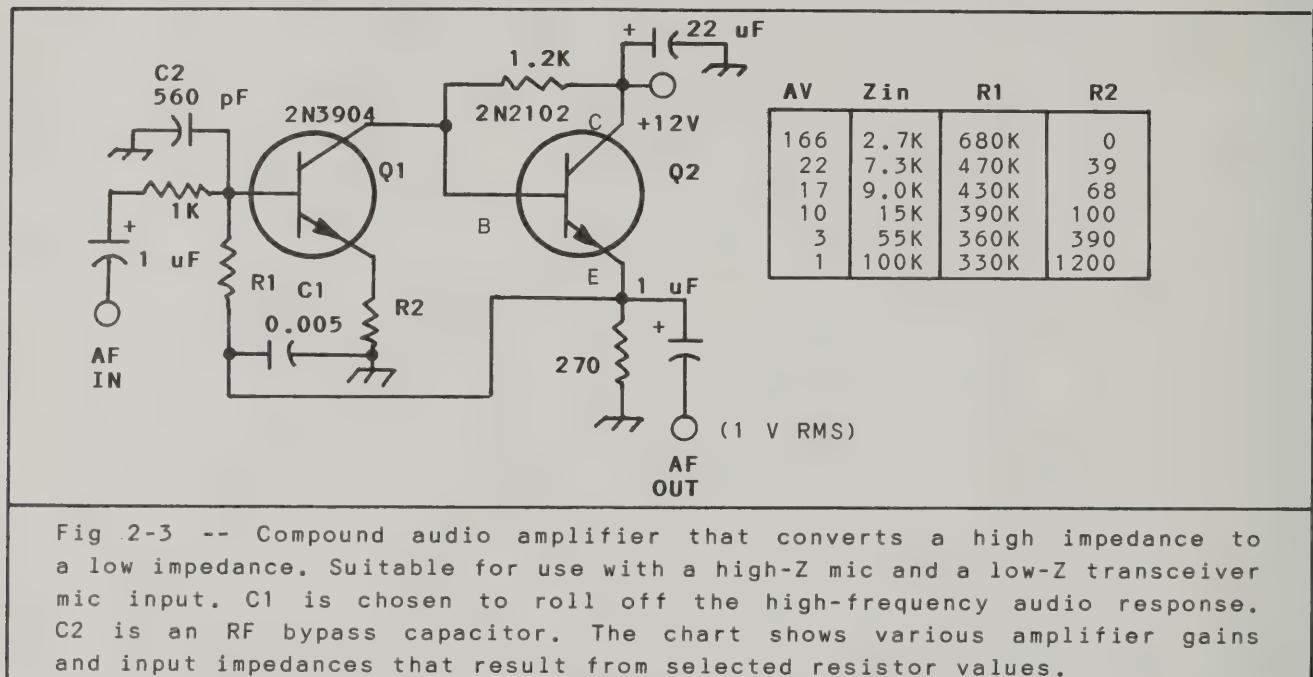


Fig 2-3 -- Compound audio amplifier that converts a high impedance to a low impedance. Suitable for use with a high-Z mic and a low-Z transceiver mic input. C1 is chosen to roll off the high-frequency audio response. C2 is an RF bypass capacitor. The chart shows various amplifier gains and input impedances that result from selected resistor values.

You may increase the value of the input and output coupling capacitors to 10 μ F if you desire greater low-frequency response. The values listed in Fig 2-3 are for rejection of frequencies below 500 Hz. In any event, do not use more than 1000 pF for C2. Larger values of capacitance will impair the high-frequency response of the amplifier. The same rule applies to C1. Use only that amount of capacitance needed to roll off the high-frequency response for the usual 500-2500 Hz speech range. You may experiment with the C1 value as outlined for the circuit in Fig 2-2.

You may use transistors other than those listed in Fig 2-3. If substitutions are made be sure to choose a device that has characteristics similar to the 2N3904 and 2N2201 transistors. The small-signal gain and power ratings of the transistors are the major consideration if you select different devices for use at Q1 and Q2.

A complete audio amplifier that drives a 4- or 8-ohm speaker is illustrated in Fig 2-4. This amplifier is suitable for use in a home-made receiver. It does not require an interstage or output transformer. Q4 and Q5 function as complimentary-symmetry transistors in the power-amplifier portion of the circuit. Note that one transistor is an NPN (Q4) and the other (Q5) is a PNP device. In order to ensure proper performance, you need to use complimentary transistors that have the same gain and power output characteristics. The 2N4401 and 2N4403 transistors provide approximately 400 mW of output. This is ample for a small receiver.

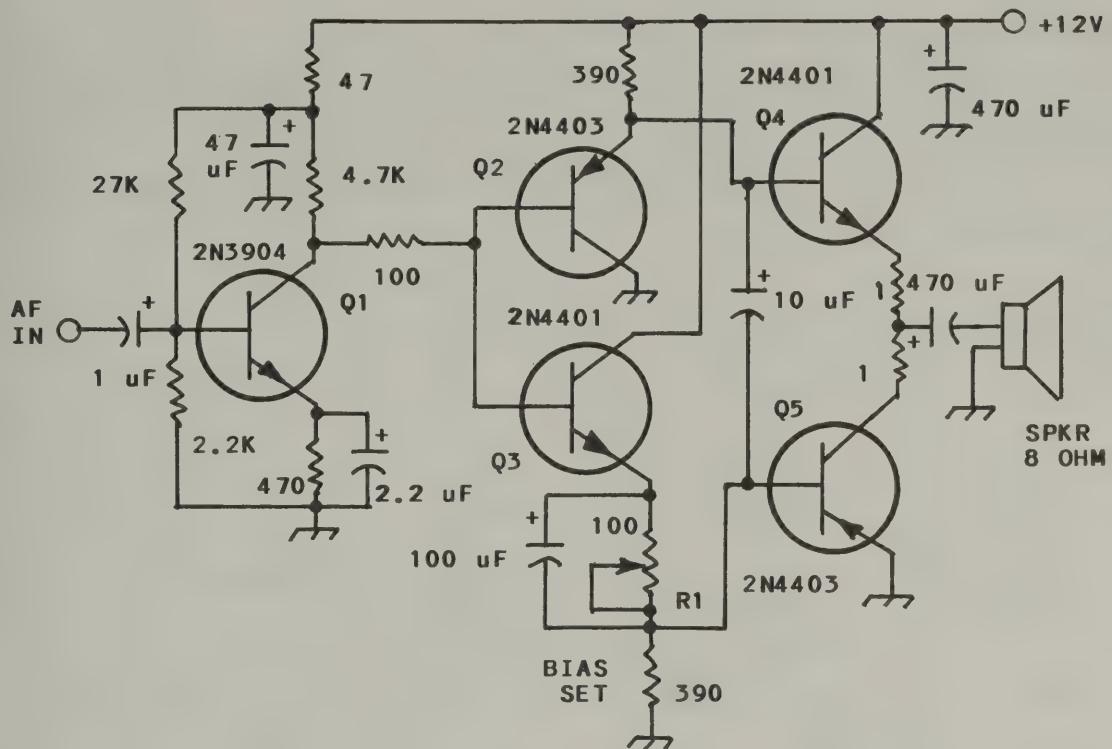


Fig 2-4 -- Schematic diagram of an audio amplifier that is suitable for use in a receiver. Power output is roughly 400 mW. R1 is adjusted for minimum audio distortion while a weak audio signal is applied to the input of the overall amplifier. Copper or aluminum tabs (3/8 X 1 inch) are epoxy-cemented to the flat sides of Q5 and Q5 to serve as heat sinks. Overall amplifier current drain is 125 mA at peak audio periods.

Various low-noise transistors may be used at Q1 of Fig 2-4. Although the 2N3904 is a common choice for low-noise audio applications, the 2N4123, MPSA09 and MPSA18 devices (among several others) are recommended by Motorola for use as low-noise audio preamplifiers. The input impedance of the Fig 4-2 amplifier is 600 ohms. You may use either a 4- or 8-ohm speaker with this circuit. The speaker should be able to accommodate 1 watt or greater audio power in order to minimize distortion. Speakers with lower power ratings may be used if you do not advance the audio gain control for volume levels beyond the rating of the speaker. A 10K-ohm audio-taper control may be used at the input of Q1 to serve as a gain control.

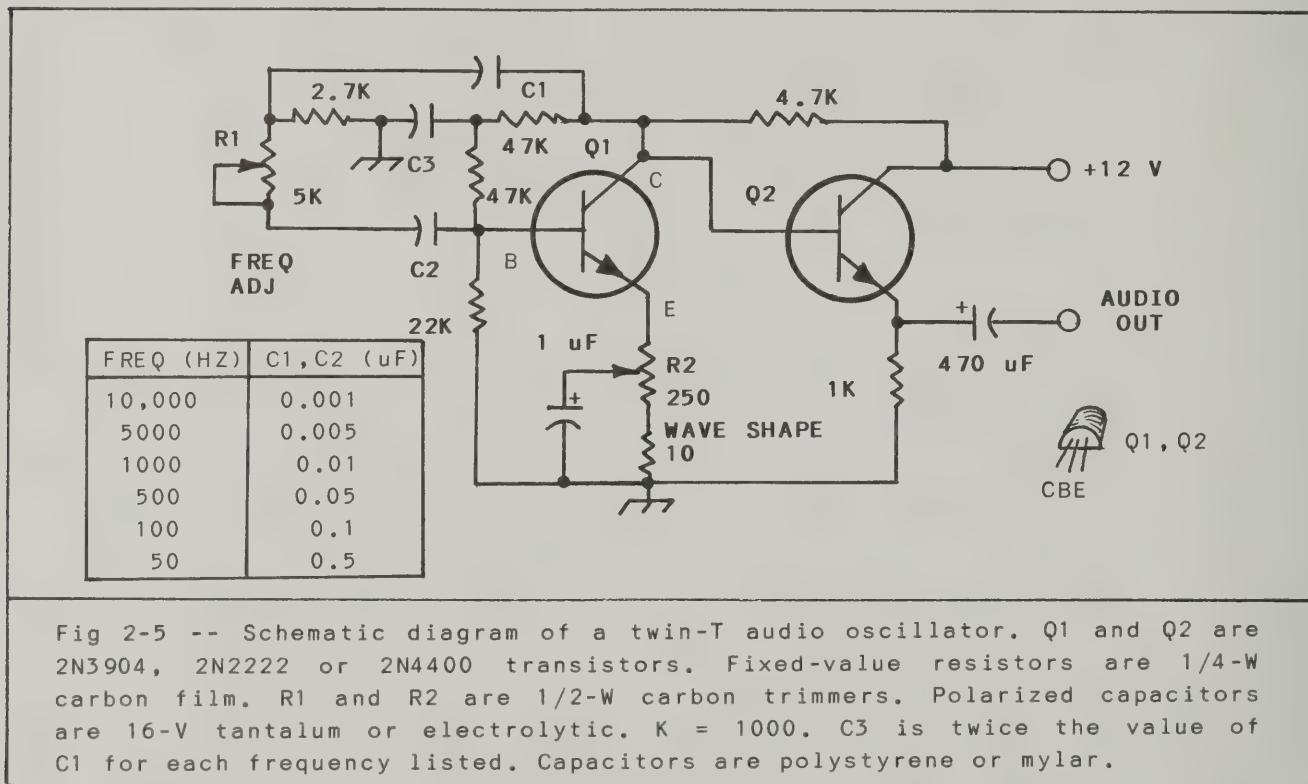
All of the polarized capacitors in Fig 2-4 are electrolytic or tantalum types. The voltage rating for these capacitors is 16 or greater. R1 is a trimmer control that can be mounted on the amplifier PC board. The 1-ohm ballast resistors for Q4 and Q5 are 1/2-W units. All other fixed-value resistors are 1/4- or 1/2-W carbon composition or carbon film units.

There are many low-cost ICs that may be used in place of the Fig 2-4 circuit, and some of them will be described in the chapter on ICs. The above amplifier can be constructed in a compact manner, should you prefer to work with discrete devices rather than ICs. If you use transistors you will have a better understanding of how the circuit operates.

OTHER AUDIO CIRCUITS

We have frequent need for an audio-frequency oscillator. We may use this circuit as a sine-wave generator that can be used when testing an audio amplifier, or we might decide to include a simple oscillator in a transmitter for use as a CW sidetone oscillator. Of course, audio oscillators are used widely as code-practice oscillators as well.

Fig 2-5 illustrates a two-transistor oscillator that was first published in the **RCA Transistor Manual**. Component values are listed in Fig 2-5 for various audio frequencies.



The above circuit has been modified from the original RCA design. R1 allows a small range of frequency change. It may be eliminated in noncritical applications. Likewise with R2 if wave shaping is not a concern. If R2 is eliminated it should be replaced by a 100-ohm resistor. Otherwise, R2 is adjusted for the purest sine wave possible, as observed with a scope. This oscillator may be keyed by breaking the +12-V supply lead. If loudspeaker operation is desired, you may feed the oscillator output into an LM386 audio IC (see section on ICs).

Mic Amplifier with Bass and Treble Boost: Many of us have voices that lack punch during SSB operation. A person's voice may lack "lows" or "highs" of sufficient amplitude to give the transmitted signal the kind of presence that causes the signal to stand out in the QRM or QRN. Some of the imported stock mics that come with transceivers have restricted frequency response, and this complicates the overall problem. Being able to boost the lows or highs, or both, can often make the poorest of mics sound acceptable. Similarly, if your voice is lacking in high frequencies, boosting the highs and rolling off the lows can solve your problem. I use the home-made W1FB electret mic that is described in August 1989

QST. It has a bass and treble boost circuit, along with a built-in audio preamplifier. Fig 2-6 contains a similar circuit that uses transistors. It has been treated for RFI suppression. It may be used with any 600-ohm mic, or it may be included in a home-made SSB or FM transmitter for use as the audio preamplifier.

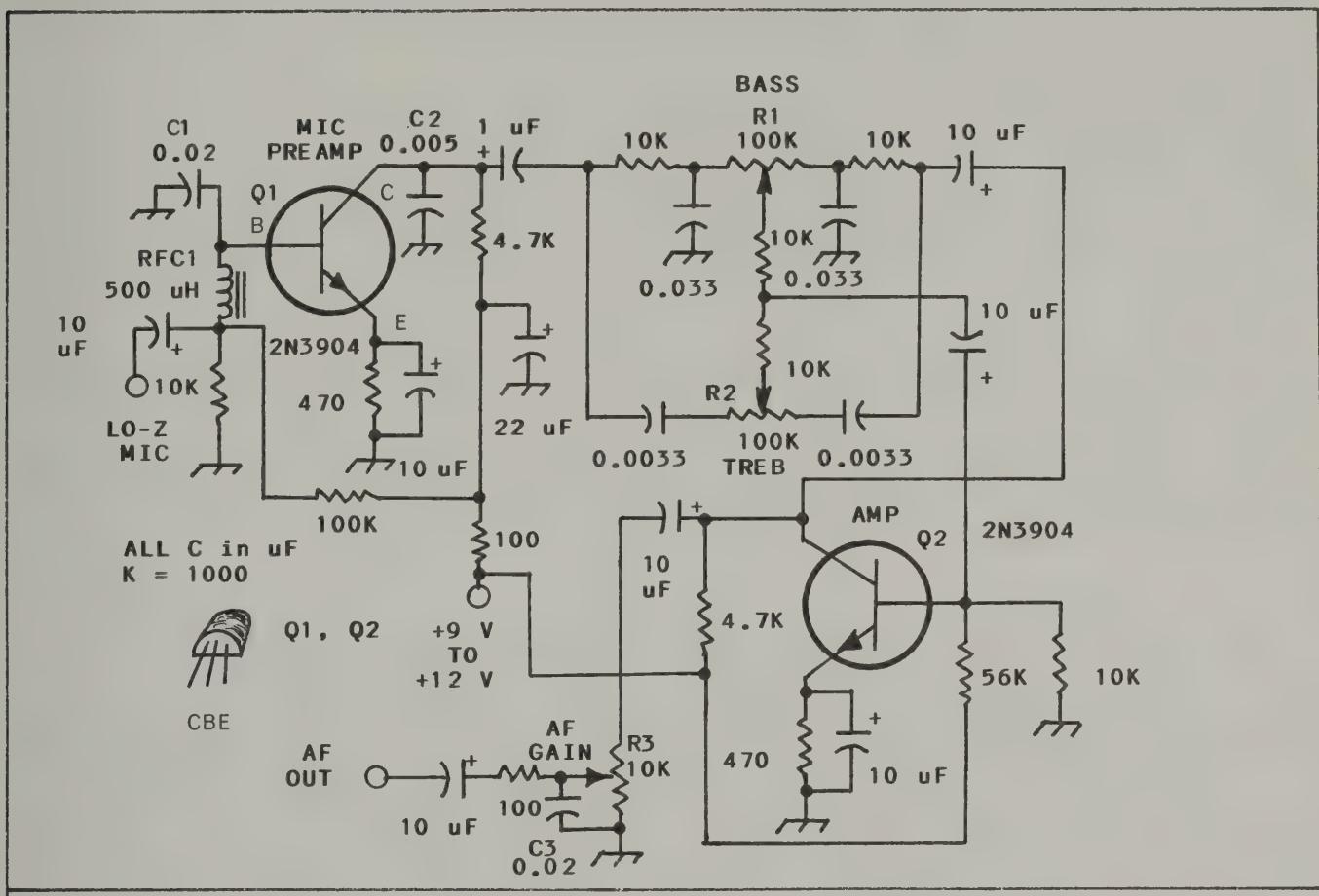


Fig 2-6 -- Schematic diagram of a bass and treble boost mic amplifier that uses the familiar Baxandall booster circuit. Fixed-value resistors are 1/4-W carbon film types. R1 and R2 are carbon PC-mount trimmers and R3 may be PC mount or panel mount carbon composition. RFC1 is a miniature 500- μ H or 1-mH choke. Polarized capacitors are tantalum or electrolytic, 16 V or greater. C1, C2 and C3 are disc ceramic. Polystyrene or mylar capacitors should be used at R1 and R2 (0.033 and 0.0033 μ F). Q1 and Q2 may be any transistor similar to the 2N3904, such as 2N2222 and 2N4400.

Capacitors C1, C2 and C3 are used for bypassing unwanted RF energy that may enter the amplifier via the mic cord or +V power lead. Too large a capacitance at these circuit points will attenuate the high-frequency response. RFC1 serves also to suppress RF energy. It should be located as close to Q1 as practicable. A ferrite bead in place of RFC1 is not effective below VHF. A single bead on a piece of wire provides approximately 1 μ H of inductance. This is not ample for filtering at HF and MF.

The Fig 2-6 circuit may be used with any low-Z mic, such as an electret or 600-ohm dynamic mic. First adjust your transmitter with your nonamplified mic for normal meter readings and a clean output signal. You may now install the Fig 2-6 booster. Adjust R3 to obtain the same meter readings during transmit. R1

and R2 may now be tweaked for the desired audio response. This can be done while using a receiver to monitor your voice transmission (wear headphones). Alternatively, you can adjust these controls while talking to someone who knows your voice characteristics. The other person can tell you when you sound natural. R1 and R2 have little effect at the voice midrange frequencies (500 to 1000 Hz), but these controls allow several dB of boost at the high- and low-frequency ends of the voice spectrum (10 to 500 and 1000 to 3000 Hz).

SMALL SIGNAL RF AMPLIFIERS

BJTs and FETs are preferred devices for low-noise RF amplification in receivers. Although some ICs may be used for RF preamplification at MF and HF, they are generally too noisy ahead of a mixer, especially at 14 MHz and higher. The exception is when we use MMICs, which have very low noise figures. MMICs are for use from VHF into the microwave spectrum, where low noise and high gain are often necessary.

A small-signal RF amplifier that is used ahead of a mixer must have a noise figure that is much lower than that of the mixer. Mixers are inherently noisy devices, and some have a negative gain (conversion loss). Therefore, the RF amplifier needs to have a low noise figure and 10 dB or greater gain in most circuits. Fig 2-7 shows circuits for common (grounded) emitter and common source RF amplifiers.

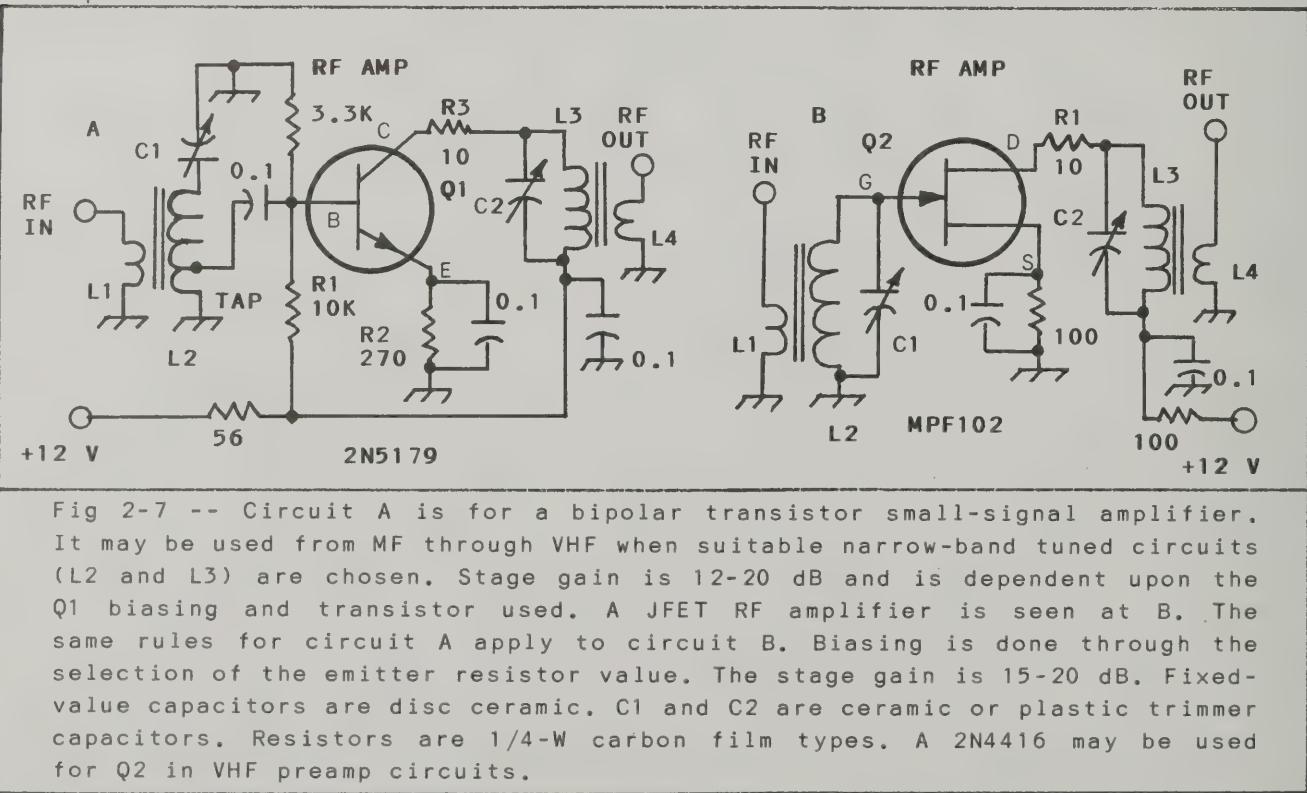


Fig 2-7 -- Circuit A is for a bipolar transistor small-signal amplifier. It may be used from MF through VHF when suitable narrow-band tuned circuits (L2 and L3) are chosen. Stage gain is 12-20 dB and is dependent upon the Q1 biasing and transistor used. A JFET RF amplifier is seen at B. The same rules for circuit A apply to circuit B. Biasing is done through the selection of the emitter resistor value. The stage gain is 15-20 dB. Fixed-value capacitors are disc ceramic. C1 and C2 are ceramic or plastic trimmer capacitors. Resistors are 1/4-W carbon film types. A 2N4416 may be used for Q2 in VHF preamp circuits.

I will offer some rules of thumb for use with the Fig 2-7 amplifiers. C1 and C2 are 300 pF maximum capacitance for 1.8 through 4 MHz, 150 pF for 4 through 14 MHz, 100 pF for 14 through 50 MHz and 50 pF for 50 through 150 MHz. Select the proper L2 and L3 inductance to provide resonance with C1 and C2 at midrange. L1 and L4 have 10% of the turns on L3 and L4, respectively. The tap on L3 of Fig 2-7A is placed at approximately 1/3 the number of turns on L2 (count up from

the grounded end of L2. If you have a means by which to measure noise figure, experiment with the Q1 biasing (R1 and R2) to obtain the lowest amplifier noise figure. This should not be necessary for frequencies below 30 MHz with the parts values and transistor specified. R3 may be eliminated if no VHF self-oscillations are encountered. This resistor acts as a parasitic suppressor. The base tap on L2 is normally selected to provide an impedance match between the input source (usually 50 ohms at L1) and the base of Q1 (typically 600-1000 ohms). Maximum stage gain will occur only if these impedances are matched.

The JFET amplifier in Fig 2-7B has fewer parts than circuit A. It is more prone to self-oscillation because both the input and output of the device exhibit a high impedance (more difficult to tame). Careful layout is necessary if you are to avoid instability. L2 and L3, along with their associated components need to be separated in order to prevent unwanted feedback caused by stray coupling. Stability in an amplifier of this type may be improved by tapping the gate toward ground on L2. The drain may also be tapped down on L3 to minimize the tendency toward self-oscillation.

Stability is enhanced when we use shielded tuned circuits. The metal shield cans are connected to circuit ground. The Amidon Assoc. L-33, L-43 and L-57 coil/transformer assemblies are excellent for this application. These components are manufactured by Micrometals Corp. and are available from Amidon with various types of cores. The core material is chosen for the operating frequency in order to ensure a suitable value of unloaded Q (Q_u). Q-versus-f data is provided in the Amidon catalog.

Toroid cores may also be used for the tuned circuits in Fig 2-7. Toroidal coils and transformers are self-shielding, and this helps to prevent unwanted inductive coupling between the tuned circuits. This minimizes the feedback that can cause self-oscillation of an amplifier stage. The toroidal tuned circuits should be spaced a reasonable distance from one another in order to avoid unwanted capacitive coupling, which may also encourage self-oscillations to occur. Various sizes and core mixes (recipes) are available from Amidon Assoc., Palomar Engineers and RadioKit when purchasing toroid cores by mail. Toroid cores are less costly than are shielded coil assemblies that feature slug tuning.

The bypass and coupling capacitors in Fig 2-7 need to have short, direct leads when they are installed. Long leads introduce unwanted stray inductance, and this inductance degrades circuit performance. At some operating frequencies the stray inductance can actually negate the desirable effects of the capacitor. Unwanted inductance may also cause an amplifier to self-oscillate. The leads on resistors need to be short and direct also in most RF circuits.

The gain of an RF amplifier needs to be tailored to the circuit needs. This is referred to generally as "gain distribution." For example, too great an RF amplifier gain, when the amplifier precedes a mixer, can seriously degrade the mixer performance (dynamic range). Too little gain, on the other hand, may not ensure that the low-noise amplifier overrides the noise of the mixer. Information on this subject is provided in ARRL's **Solid State Design for the Radio Amateur**. Gain distribution is an important consideration in any circuit that uses many active stages. In transmitters, for example, too little stage gain will result in low transmitter output power. Too much gain can overdrive a succeeding RF amplifier stage and cause it to saturate, draw excessive current and possibly burn out. Excessive driving power also enriches harmonic currents.

Other FET RF Amplifiers

You may use dual-gate MOSFETs as RF small-signal RF amplifiers. Fig 2-8 shows a 3N211 MOSFET as a RF amplifier. The signal is applied to gate no. 1 and positive bias is supplied to gate no. 2 in circuit A. The JFET in Fig 2-8B is operated in the common-gate or grounded-gate mode. This circuit is very stable if the gate lead is kept very short between the device and ground. The input impedance is roughly 200 ohms ($Z_{in} = 10^6/gm$). The gain of a common-gate amplifier is somewhat less than that of a common-source amplifier. Typical gain for a common-gate amplifier is 10-12 dB, whereas you may expect up to 20 dB of gain from a well designed common-source amplifier.

Dual-gate MOSFETs may be used as single-gate transistors by connecting gate 1 and gate 2 together and treating them as a single gate. In this manner they may be substituted for a JFET in most circuits. No gate bias is required when this is done.

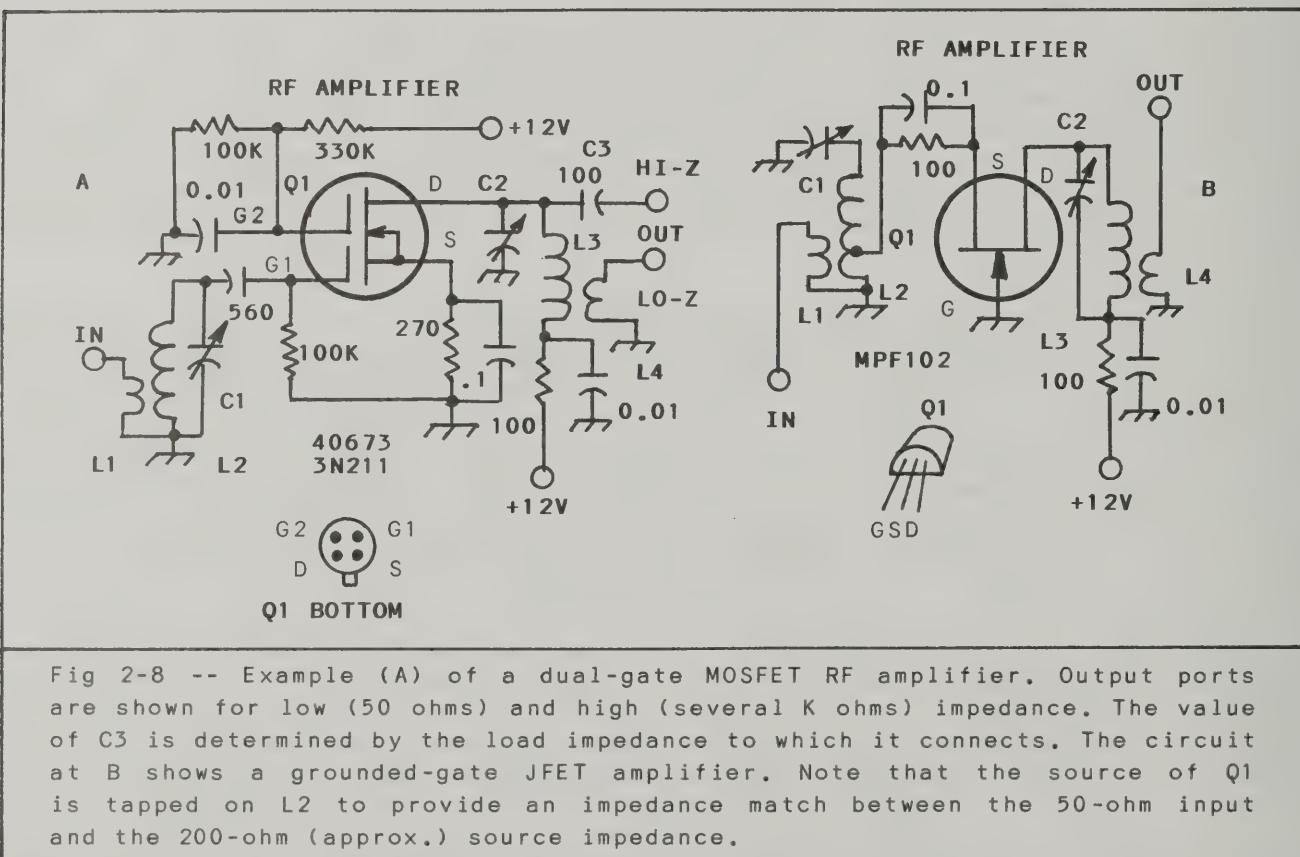


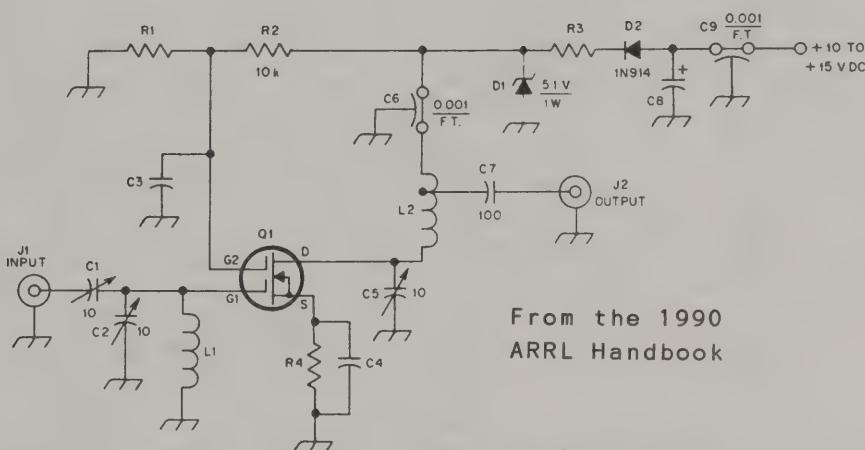
Fig 2-8 -- Example (A) of a dual-gate MOSFET RF amplifier. Output ports are shown for low (50 ohms) and high (several K ohms) impedance. The value of C3 is determined by the load impedance to which it connects. The circuit at B shows a grounded-gate JFET amplifier. Note that the source of Q1 is tapped on L2 to provide an impedance match between the 50-ohm input and the 200-ohm (approx.) source impedance.

The RF amplifier in Fig 2-8A can provide up to 20 dB of gain. High- and low-impedance outputs are shown. The lower the load impedance connected to C3 the smaller the capacitor value. This prevents excessive tuned-circuit loading, which can destroy the tuned-circuit Q. For example, if C3 is connected to a 50-ohm load, its value should be less than 20 pF from 1.8 to 10 MHz. If the output load is, say, 2000 ohms, C3 might be increased to 47 pF. C1, C2, L2 and L3 are chosen for the operating frequency. The LC ratio is not critical. L1 and L4, by rule of thumb, contain approximately 10 percent of the total L2 and L3 coil turns for a 50-ohm output impedance. The number of L1 and L4 turns may be determined experimentally for best circuit performance.

The common-gate JFET amplifier in Fig 2-8B is less complicated than that shown at A of the same figure. The maximum available amplifier gain is lower than for the dual-gate MOSFET example. Typical gain for the circuit at B is 10-12 dB. Both circuits offer low-noise performance from MF into the VHF spectrum. The source of Q1 is tapped on L2 to provide a match between the input and source impedance. A suitable experimental location for the tap is approximately 1/4 the total L2 turns (count up from the grounded end of L2). The tap point may be changed to obtain maximum stage gain, consistent with proper C1/L2 selectivity. Too high a tap point will load the tuned circuit and lower its Q. You may use other JFETs for Q1 of Fig 2-8B, such as the 2N4416 and other HF/VHF devices.

GaAsFET Amplifiers

Gallium arsenide FETs are well suited to low-noise operation at VHF and above. Their performance in this regard far exceeds that of 3N211s and MPF102s. Fig 2-9 illustrates a VHF amplifier that uses a GaAsFET as the active device.



From the 1990
ARRL Handbook

Fig 2-9 -- Circuit for a GaAsFET 432-MHz amplifier. Q1 is a dual-gate FET (NEC NE41137, Motorola MRF966/967 or a 3SK97). Fixed-value capacitors are ceramic chip or leadless units to provide a low-Z path for the RF energy. Complete parts-value data is provided in **The ARRL Handbook** (1990) on pages 32-1 and 32-2. Noise figure is 0.7 to 0.9 dB and the amplifier gain is 22 to 24 dB.

GaAsFETs are available also in the JFET configuration, such as the Mitsubishi MGF1402. A 432-MHz preamplifier that uses this device is also shown in the 1990 **ARRL Handbook**, page 32-2, along with GaAsFET circuits for 1296 MHz. The resistor values in Fig 2-9 are selected for the GaAsFET used at Q1. This information is given in **The ARRL Handbook**.

Bipolar transistors are used also for VHF, UHF and microwave RF amplifiers. The MRF901, for example, is capable of a noise figure less than 2.5 dB at 1296 MHz. Power gains up to 10 dB are common when using the MRF901 at 1296 MHz. A cost-effective 1296-MHz preamp with an MRF901 may be seen on page 32-9 of the 1990 **ARRL Handbook**.

Broadband RF Amplifiers

The RF amplifiers we have discussed thus far are for narrow-band applications (tuned circuits). Broadband amplifiers have no selectivity, per se, but they

are essential when we need an amplifier that covers several MHz without being adjusted to each new operating frequency. Broadband amplifiers require feedback circuits to ensure that the gain is reasonably flat (constant) across the frequency range of the amplifier, such as 1.8 to 30 MHz. Broadband amplifiers are generally designed to operate linearly, and they have predictable input and output impedances. This makes them ideal for the designer who needs a broadband amplifier for AM or SSB signal amplification. In order to enhance the broadband characteristics of this amplifier we must use broadband inductors or transformers in place of the tuned circuits shown in Figs 2-7, 2-8 and 2-9. Broadband amplifiers are inherently more noisy than the circuits shown earlier in this section. Therefore, they are seldom used as low-noise preamplifiers for receivers. Rather, they find common use as IF (intermediate frequency) amplifiers and RF power amplifiers. Fig 2-10 shows a practical low-power broadband RF amplifier. Broadband power amplifiers will be discussed later.

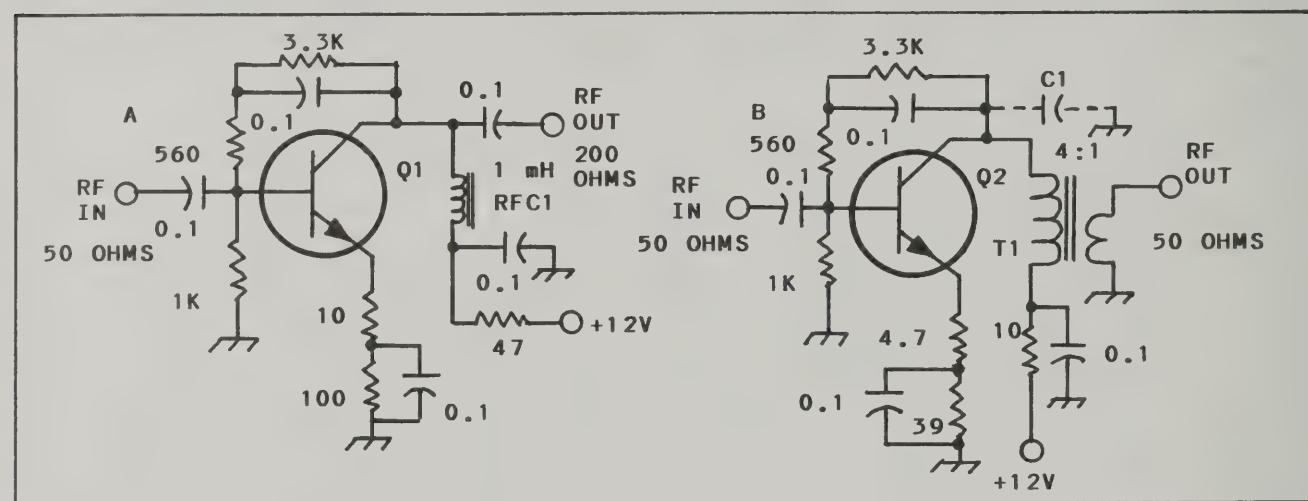


Fig 2-10 -- Examples of broadband amplifiers for use from 1 to 50 MHz. Circuit A is for small-signal linear amplification. Q1 is a 2N5179 for efficient use at the upper end of the HF and lower part of the VHF spectrum. A 2N5770 may be used in place of the 2N5179. Other devices, such as the 2N2222A and 2N4401 may be used at A for operation below 30 MHz. The circuit at B is capable of greater output power than the example at A. It uses a huskier transistor, such as the 2N5109 or 2N3866. T1 provides a step down from 200 to 50 ohms. T1 has 16 primary turns of no. 26 enam. wire on an Amidon FT-50-43 (850 mu) ferrite toroid. The T1 secondary has 8 turns of no. 26 wire. A 4:1 transmission-line transformer may be used also at T1. RFC1 in circuit A may be replaced by T1 in circuit B if a 50-ohm output is desired.

The circuits in Fig 2-10 use shunt feedback between the base and collector. Degenerative feedback (unbypassed emitter resistors) is used also. It helps to provide a 50-ohm input impedance for Q1 and Q2. The values of the bypassed emitter resistors (39 and 100 ohm) can be chosen to establish the desired no-signal (quiescent) collector current of the transistor.

C1 in Fig 2-10B may be included for bypassing VHF harmonics in the amplifier output. The XL of C1 should be no less than 800 ohms at the upper operating frequency in order to minimize signal loss in the HF range of the amplifier. Thus, for a 1.8 to 30 MHz broadband amplifier we will use nothing larger than 6.6 pF. This becomes more practical when the broadband amplifier is for use in a single, lower-frequency band such as 3.5-4.0 MHz. In this example we use

4 MHz as the upper frequency. This results in a capacitance of 49 pF for an X_L of 800, which is 4×200 ohms. C_1 should not be required except in the most critical of amplifier circuits.

We should use transistors that have a high f_T and beta when building broadband amplifiers. The 2N5179, for example, has an f_T of 1200 MHz and a small-signal beta (h_{fe}) of up to 300. Q_2 in Fig 2-10B is adjusted for a no-signal collector current of approximately 50 mA. A press-on crown heat sink is recommended for keeping the transistor cool.

The amplifiers in Fig 2-10 provide a gain of 10-15 dB when matched correctly to their input and output loads. They should be unconditionally stable if the circuit layout is done carefully. The huskier amplifier at B is suitable for use as a driver in a transmitter, or for use as a post-mixer amplifier in a receiver that features high dynamic range.

RF Power Amplifiers

Bipolar transistors and power FETs are used for the generation of high levels of RF power. The most common classes of amplifier operation for amateur work are Class A and Class C. A narrow-band Class C amplifier is capable of an efficiency as great as 70% for bipolar devices, and up to 85% for power FETs. Class AB amplifiers are used in the broadband mode, and they exhibit efficiencies in the 50-60 percent range. Likewise for broadband Class C amplifiers. There must always be a tradeoff between efficiency and bandwidth. The advantage we realize during broadband operation is that band switching is simplified and no tuning is required. It becomes necessary only to switch a group of 50-ohm harmonic filters when using a multiband amplifier that is broadbanded.

Class A amplifiers are biased for linear operation. This makes them suitable for AM and SSB amplification. Class C amplifiers may be used in the output of an AM transmitter if modulation is applied to the amplifier. They are used also for the amplification of CW and FM signals, where linearity is not required. Linear amplifiers may, however, be used for amplifying CW and FM signals. Simple examples of RF power amplifiers are presented in Fig 2-11.

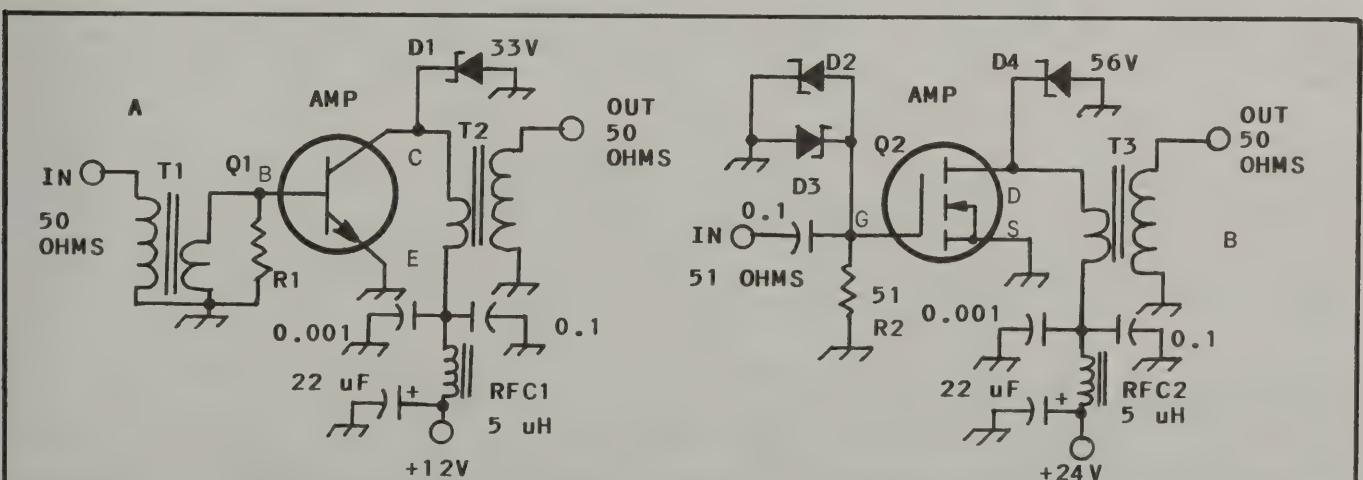


Fig 2-11 -- Example A shows a single-ended Class C BJT amplifier. The circuit at B illustrates a Class C power MOSFET broadband amplifier (see text for further information).

The simple amplifier in Fig 2-11A exhibits a relatively low input impedance. This is usually between 1 and 10 ohms, depending on the transistor used and the current flowing through the base-emitter junction. T1 is wound to provide an impedance match between 50 ohms and the Q1 base. A ball-park impedance ratio of 5:1 for T1 represents a suitable starting point. R1 is often used in this type of circuit to aid amplifier stability. Values from 10 to 47 ohms are common for the MF and HF spectrum. Use the highest value that prevents amplifier self-oscillation. The lower the R1 value the greater the input-power loss.

The output of the circuit in Fig 2-11A has a broadband transformer for T2, as is the situation for T1. An impedance match between the 50-ohm amplifier load and the Q1 collector is effected by means of T2. The collector impedance is found from V_{cc}^2 divided by $2 P_o$, where V_{cc} is the supply voltage and P_o is the output power in watts. The transformer turns ratio is adjusted accordingly. The 50-ohm T2 winding may be connected to a 50-ohm low-pass harmonic filter. Zener diode D1 may be added to protect Q1 from voltage spikes that may appear on the +12-V supply line. The diode also protects the transistor when the SWR is high or if the amplifier should break into self-oscillation. In other words, D1 limits the collector-voltage swing to +33 V. A 1-W Zener diode is recommended.

RFC1 and the associated bypass capacitors provide bypassing from audio through the VHF spectrum. The choke may consist of a few turns of no. 20 enamel wire on an FT-50-43 ferrite toroid. The wire must be able to carry the Q1 collector current without causing a voltage drop. The T1 and T2 toroid or balun cores are also of ferrite, and have a permeability of 800 to 1000 for HF operation. Amidon or Fair-Rite no. 43 material is suitable.

A power MOSFET amplifier is illustrated at B of Fig 2-11. Note that R2 is used for establishing the amplifier input impedance. A power FET has a gate impedance of 1 megohm or greater without R2. These enhancement-mode FETs require a peak-to-peak gate voltage of up to 30 in order to turn on and produce output power. Therefore, the lower the R2 value the greater the driving power needed to develop the desired voltage swing across R2. Values as great as 1000 ohms have been used for R2, provided an input matching transformer was included.

D2 and D3 may be added to protect the fragile FET gate from excessive voltage and current. A pair of 15-V, 400-mW Zener diodes are suitable. The diodes do, however, add shunt input capacitance to the circuit. This can complicate the impedance match, especially at the upper end of the HF spectrum. A protective Zener diode (D4) may be connected from the Q2 drain to ground for the reasons given during the discussion of Fig 2-11A. Some power FETs have a built-in drain-source Zener diode. In this situation you may eliminate D4. The drain impedance is determined by $Z(\text{ohms}) = V_{dd}^2$ divided by $2P_o$. This equation provides an approximate Z value. T3 is wound to match the 50-ohm load to the Q2 drain. A suitable harmonic filter is used at the output port of T3.

VHF parasitic oscillation is a common malady when working with power FETs. A quick solution for the problem is to install a 10- or 15-ohm carbon composition resistor (1/2 W) in series with the signal path. It should be located as close to the transistor gate as practicable. A ferrite bead may be used in place of the resistor for de-Qing the input circuit.

Class A linear operation of the Fig 2-11 circuits is shown in Fig 2-12. You will notice that the circuits are similar. The only change is related to the

application of forward (positive) bias to the base of the BJT and the gate of the power FET.

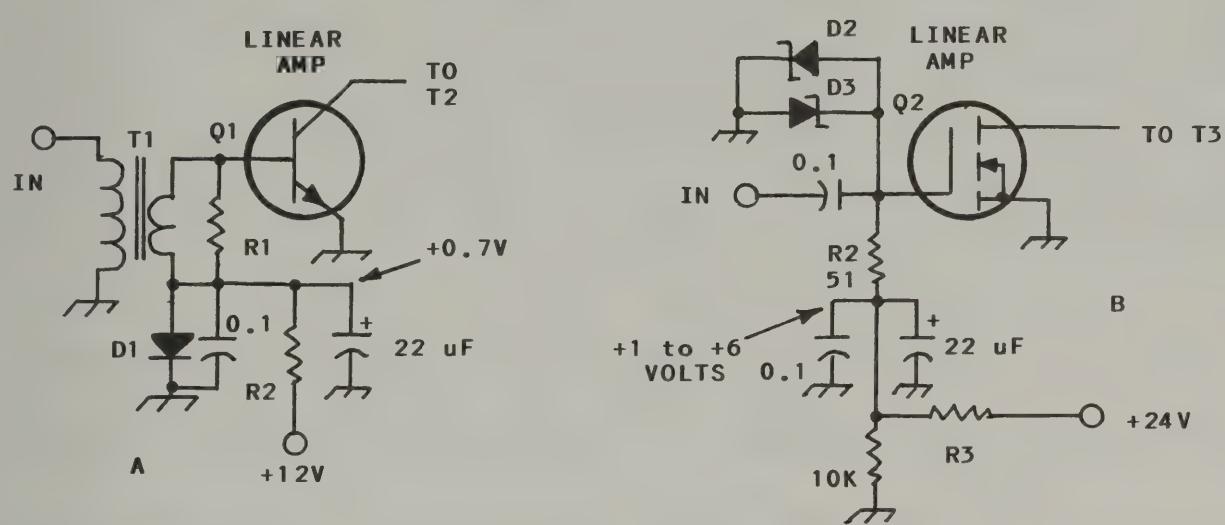


Fig 2-12 -- Examples of linear amplifiers as derived from the circuits in Fig 2-11. D1 at A is a rectifier type of diode, 50 PRV at 1 or 2 A. Its barrier voltage (0.7) establishes a regulated forward bias for Q1. R2 is selected to permit 50-100 mA to flow through the D1 junction. A 2-W resistor is required. Circuit B obtains its bias from a simple resistive divider. The Q2 gate current is in uA, so 1/4-W resistors may be used in the divider. R3 is selected to provide the desired forward bias for Q2. A no-signal drain current of 100 mA or greater is typical for Class A operation. The higher the forward bias the greater the drain current.

Simple bias circuits are shown in Fig 2-12. Some commercial designers use IC voltage regulators and variable bias-voltage controls. In any event it is important that the forward bias for a linear amplifier be stiff (well regulated) if linearity is to be preserved. The 22-uF bypass capacitors on the bias lines help to stabilize the voltage by maintaining a dc charge at the bias-voltage value. Momentary variations in the Vcc or Vdd supply line are masked by the presence of these capacitors.

It is important that solid-state linear amplifiers have heat sinks large enough to keep the transistor junction temperature at a safe level. Linear amplifiers require a larger heat-sink area than do Class C amplifiers that operate in the same power range. This is because the duty cycle is more severe in a linear amplifier, plus the idling current that keeps the devices hot. No power transistor should be operated at a case temperature greater than the spec sheet states.

Power FETs are not subject to thermal runaway, and they are relatively tolerant of SWR. They are, however, more fragile than BJTs with regard to voltage spikes and excessive gate current. The thin layer of silicon oxide that insulates the FET gate from the drain-source junction can be perforated quickly by a voltage spike. These spikes may appear on the supply voltage, and may occur on the RF input voltage line as well. Linear amplifiers generate less harmonic energy than is found in the output of Class C amplifiers. High-order IMD (intermod distortion products) are of lower amplitude in a linear FET amplifier than they

are in a linear BJT amplifier -- a definite plus feature.

Power FETs that are designed specifically for RF service can deliver their full rated output to approximately 175 MHz. Although power FETs that are designed for ac and dc switching, such as the IRF511, can be made to work quite well as RF amplifiers up to roughly 14 MHz, they exhibit poor efficiency at the upper end of the HF spectrum, and they are useless at VHF. This is because the internal geometry (the HEXFET is an example) is large, and the internal capacitances are much higher than in an RF type of power FET. The V-groove technology, such as that developed by Siliconix Corp., provides a much better RF power FET. The gate impedance of a switching FET is composed primarily of XC (capacitive reactance) and this establishes a "frequency sensitivity" that limits the practical upper frequency of the transistor. Conversely, the input capacitance of a BJT becomes greater as the operating frequency in MHz is lowered. This, in a sense, tends to correct the frequency-sensitivity problem. Power FETs, on the other hand, maintain a fairly constant input and output capacitance versus frequency. This makes the FET an excellent device for use in a linear amplifier that has shunt feedback: the feedback network is much easier to design than is the case when designing the feedback network for a wide-band BJT RF power amplifier.

Although I have shown single-ended RF power amplifiers in this section, I want to acknowledge the practicality of amplifiers that use transistors in push-pull or parallel. Push-pull operation is superior to parallel operation when we wish to increase the amplifier power over a single device. Push-pull operation results in a lower power level for even-order harmonics. The output impedance of the transistors is always higher in a push-pull amplifier than in a circuit that has the same two transistors in parallel. We often encounter collector impedances (only a few ohms) in parallel operation that are impractical to match with a broadband transformer.

Several individual push-pull amplifiers may be used in concert to develop very high amounts of RF output power. This is done by using hybrid splitters at the inputs of the various identical amplifiers. This divides the driving power among the amplifiers in an equal proportion. The amplified outputs are joined by means of hybrid combiners. Splitters and combiners are broadband transformers with windings that have a specific phase relationship. Combiners and splitters are discussed (with practical examples) in **Solid State Design for the Radio Amateur**, and in several of the Motorola application notes by Helge Granberg, K7ES.

TRANSISTOR OSCILLATORS

We can think of oscillators as tiny transmitters. In fact, amateurs have for many decades used oscillators as low-power transmitters. The present QRP movement is tailor made for single-transistor oscillators. An oscillator is not as efficient as an amplifier, and this is because some of the oscillator output power is routed back to the input circuit as feedback energy. An oscillator is actually an amplifier, and the feedback is used to make it oscillate. The feedback energy (about 25% of the output power) is expended to provide oscillation. A typical oscillator has an efficiency of less than 50 percent.

Too little feedback can prevent oscillation or make an oscillator slow to start. Too great a feedback amount can cause multiple output frequencies, damage a crystal or cause even a crystal-controlled oscillator to exhibit frequency drift. We should adjust the feedback so that we use only the amount that assures rapid oscillator starting when the operating voltage is applied. The feedback should

be consistent with minimal long-term frequency drift.

Feedback also has an effect on the purity of the oscillator output waveform. Oscillator biasing affects the output wave shape too. An ideal oscillator would have a pure sine wave at its output port, but this is seldom the case because of limiting and harmonic currents.

Output coupling for an oscillator should be light in order to prevent excessive loading by circuits that follow the oscillator. Excessive coupling can prevent an oscillator from starting when power is applied. If the output coupling is too great it can lead to frequency shifting as the output load is keyed or tuned. Load changes cause phase shifts, and these affect the operating frequency. This light coupling, though necessary, also contributes to a reduction in the effective oscillator efficiency.

Oscillators are usually isolated from their loads by means of one, two or three buffer amplifiers. The better the isolation between the oscillator and its load, the less pronounced the frequency shift or "pulling" when the load shifts or is applied. This is more important when using VFOs than when operating a crystal-controlled oscillator. We should be aware, however, that fundamental and overtone crystal oscillators that operate in the upper HF and the VHF spectrum may be affected significantly by load changes. Buffer stages may be required with these oscillators as well. Transistors, compared to vacuum tubes, are more subject to pulling because their internal capacitances are substantially greater than those of vacuum tubes.

Crystal-Controlled Oscillators

The most common crystal oscillators are the Pierce, Colpitts and standard overtone types. A host of other oscillator circuits exist, and most of them are named after the person who developed them. Our concern in this section is focused on the most common of the circuits and how they are configured. Fig 2-13 contains practical circuits for these oscillators.

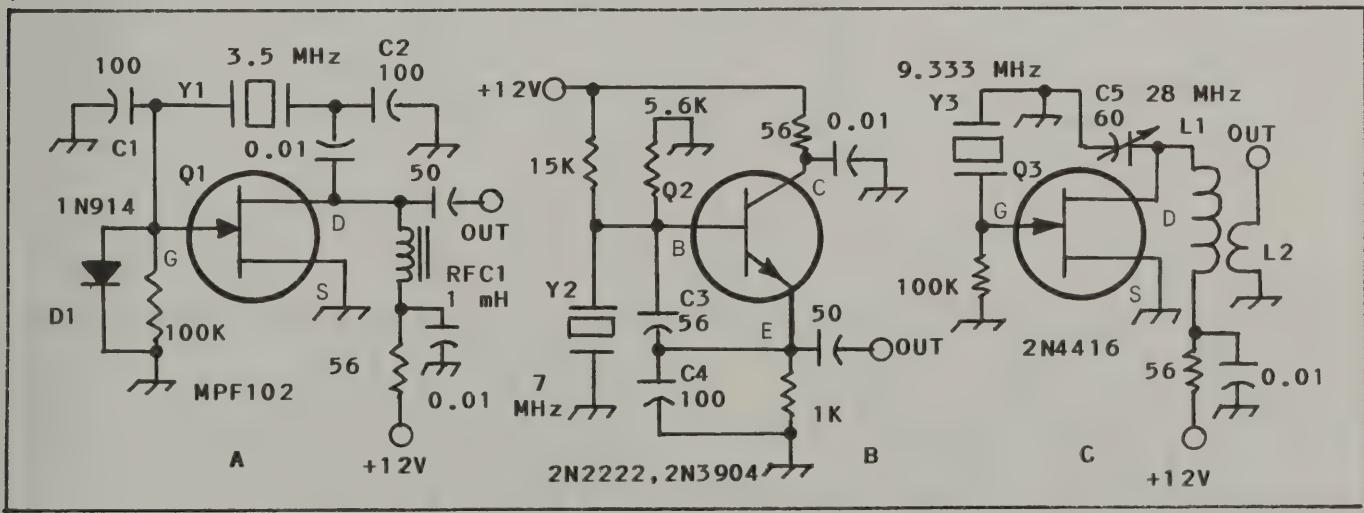
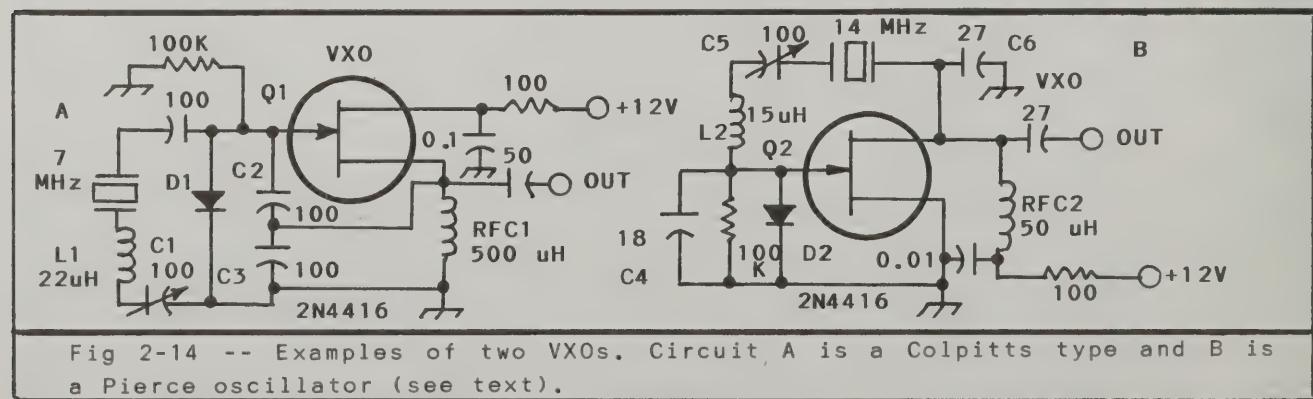


Fig 2-13 -- Practical examples of crystal oscillators. Circuit A is a Pierce oscillator. A BJT may be used at Q1 if the biasing shown at B is used. Circuit B is a Colpitts oscillator for which C3 and C4 determine the level of feedback. An overtone oscillator is shown at C. L1 is tuned to the desired odd harmonic of the crystal, 3rd, 5th, 7th, etc.

The Pierce oscillator in Fig 2-13A depends upon C1 and C2 for feedback control. The feedback path is through Y1, from the Q1 drain to its gate. C1 and C2 are chosen to provide only the feedback needed to sustain oscillation. The capacitor values are sometimes equal, as shown, but C2 is frequently much larger than C1. This depends on the crystal activity and the transconductance of the FET. Values as small as 15 pF may be used at C1 and values as large as 1000 pF can be used at C2. The lower the oscillator operating frequency the larger the C1 and C2 values, generally speaking. RFC1 and the related stray capacitance must resonate 1 MHz or greater below the frequency of Y1 in order for this circuit to function. Do not use a tuned circuit in the output of Q1 for Pierce operation. A tuned circuit may be added if you wish to use the Fig 2-13A circuit as an overtone oscillator. In this case you will eliminate C1 and C2. A tuned drain circuit is added and adjusted to the desired odd harmonic of the crystal. You may use a trimmer capacitor in this event. RFC1 is replaced by a coil (as in Fig 2-13C). A bipolar transistor may be used in place of a JFET at Q1. If this is done, the base must be biased, as in the circuit of Fig 2-13B.

Fig 2-13B shows a Colpitts crystal oscillator. A JFET may be used at Q2 by grounding the source directly and removing the forward bias resistors. C3 and C4 set the feedback amount. Their values are increased as the operating frequency is lowered. In a similar fashion, their values are decreased as the operating frequency is increased. C3 is chosen to provide the desired amount of feedback. The higher the FET transconductance and the greater the hfe of a BJT, the better the device is for oscillator service. The fT of the BJT should be at least five times the operating frequency for best performance, even though most transistors will oscillate at their fT or even slightly above it.

A 3rd-overtone oscillator is seen at C of Fig 2-13. Here again we may use a BJT rather than a JFET. C5 and L1 tune the output to the desired crystal overtone frequency. When this is accomplished the crystal oscillates only on its chosen overtone. There should be no fundamental energy present. The efficiency of an overtone oscillator is lower than that of a fundamental crystal oscillator. Although most fundamental crystals can be made to oscillate on their overtones, it is better to use a crystal that is designed for overtone use. Be aware also that the overtone frequency is not necessarily an exact multiple of the fundamental crystal frequency. It is sometimes desirable to use regulated voltage to operate an overtone oscillator. Frequency drift can occur with crystals that are used at the upper HF range. The drift can become rather significant at the overtone frequency. Slight shifts in operating voltage can result in an attendant change in oscillator frequency. Two VXOs (variable crystal oscillator) are depicted in Fig 2-14.



A VXO is substantially more stable than an LC type of VFO. The shortcoming of VXOs is that they provide a limited frequency change. The lower the operating frequency the smaller the frequency change. For example, a 3.5-MHz VXO does not provide more than approximately 1.5 kHz of frequency shift. A 21-MHz VXO, on the other hand, can be shifted some 20 kHz with ease. The circuit in Fig 2-14A (7 MHz) will allow a shift of roughly 8 kHz with a plated, AT-cut crystal. If the inductance of L1 is increased by a large amount, the oscillator can be shifted 100 kHz or greater, but the stability becomes degraded. The circuit actually operates as an LC type of VFO under these conditions. C2 and C3 are the feedback capacitors. C1 is used to tune the available range of the VXO. VXO frequency change is nonlinear, with maximum change occurring toward minimum capacitance of C1. L1 may be a miniature RF choke. Bipolar transistors may be used in place of the JFETs shown in Fig 2-14. The addition of base and emitter bias resistors is required when using BJTs at Q1 and Q2.

Fig 2-14B shows a Pierce type of VXO. C4 and C6 are the feedback capacitors. C5 must have its rotor and stator insulated from ground in this circuit. With the circuit values shown, the VXO will yield a frequency shift of approximately 15 kHz.

The upper operating frequencies of the xtals in Fig 2-14 will be roughly 2 kHz higher than the marked values when C1 and C5 are at minimum capacitance. As the tuning capacitors are adjusted for maximum capacitance, the operating frequency becomes lower. The frequency change becomes slower and slower as maximum capacitance is reached.

Standard Variable-Frequency Oscillators

The principles of coil-capacitor oscillators (VFOs) are similar to those for crystal-controlled oscillators. Feedback energy is required to make the basic amplifier self-oscillate, and frequency stability is a major concern. In fact, short- and long-term stability is one of the biggest challenges to a ham designer of VFOs. The higher the VFO operating frequency in MHz the greater the care needed to ensure acceptable stability. VFOs that operate on fundamental frequencies above, say, 10 MHz are generally impractical for use in communications circuits that have receivers with narrow filters. It is better to operate the VFO at a lower frequency and include multiplier stages to reach the desired operating frequency. Heterodyne VFOs are the better choice, although they are somewhat more complex than direct VFOs.

Semiconductors, as opposed to vacuum tubes, are more prone to frequency drift when used in VFOs. This is because transistors have internal capacitances that change with heat. This heat is caused by both dc and RF current flow through the transistor junction. In a like manner, the internal capacitances of a solid-state buffer amplifier will change with variations in operating temperature, and this prevents the oscillator from having a constant load. Load changes, as we learned earlier, cause frequency changes.

In order to minimize the effects of dc and RF heating we can operate the VFO at reduced, regulated voltage. For example, we may elect to use only +6 V for the oscillator part of a VFO chain. This reduces the circulating current, but reduces the available VFO output power. A post-oscillator amplifier can be used to elevate the effective output power of the VFO.

It is essential that we use high-Q coils and capacitors in our VFOs. The greater the Q the better the performance and the lower the wide-band noise output. Wide-band noise may appear on a transmitted signal and it can seriously degrade the performance of a receiver. An ideal oscillator would have zero noise output.

VFO coils are the most stable when they are air-wound and rigid. The introduction of magnetic core materials in a coil, such as powdered-iron toroids and slugs, or ferrite toroids or slugs, has a marked effect on oscillator stability. These core materials undergo permeability changes with variations in ambient temperature. In some instances they undergo small changes in size, and this affects the coil inductance. Heat can alter the dimensions of magnetic cores, however minuscule the changes may be. If core material is used, try to adhere to no. 6 powdered iron, since it has excellent stability. Also, it yields a high coil Q up to approximately 50 MHz. After the toroidal coil is wound it should be doped with two coatings of polystrene Q Dope. This will keep the coil turns from shifting and causing frequency changes.

Fixed-value VFO capacitors play a vital role in frequency stability. I strongly recommend zero temperature coefficient (NPO) ceramic capacitors in the frequency-determining part of a VFO circuit. This includes the feedback and output-coupling capacitors. My next choice is polystyrene capacitors, which are quite temperature-stable too. Silver-mica capacitors are third on the list. They are rather unpredictable with regard to stability versus heating. Some exhibit a positive drift factor and some have a negative factor, even if they are from a given production run. It often requires some cut-and-try effort to find a group of silver-mica capacitors that provide stable VFO operation. Polystyrene capacitors have a negative drift trait, whereas magnetic cores have a positive-drift characteristic. These two components, in combination, offer stable operation because their drift factors tend to cancel one another.

Avoid the use of double-sided PC boards for your VFOs. The PC elements form small capacitors with the ground-plane side of these boards, and capacitors that are formed in this manner are extremely unstable. Try to keep your overall VFO circuit in a separate shielded compartment. This prevents stray RF energy from entering the VFO circuit. Stray RF currents cause instability and erratic operation.

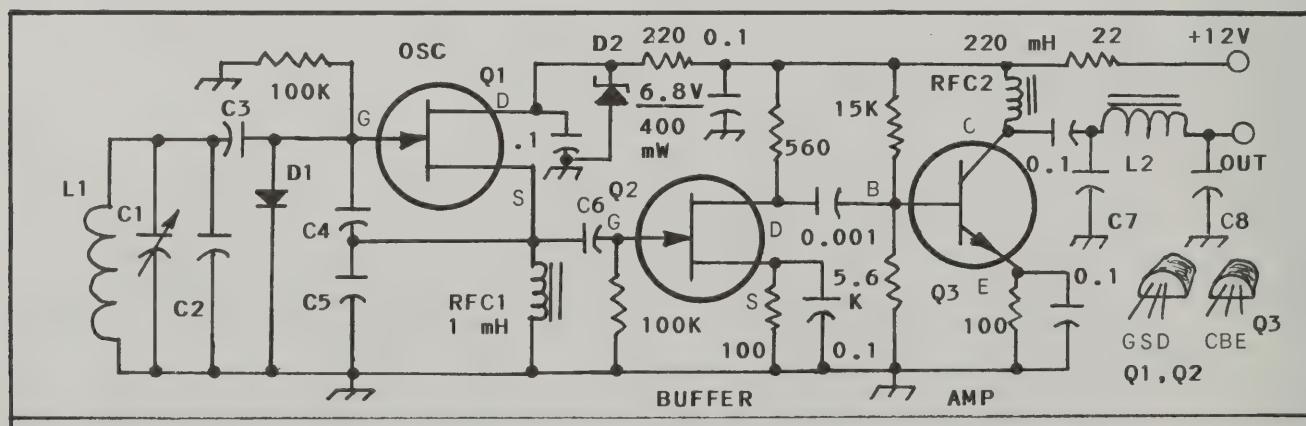


Fig 2-15 -- Basic circuit for a VFO with a buffer and an amplifier. C2-C6 are temperature-stable NPO or polystyrene capacitors. C1 is the main tuning control. Q1 and Q2 are MPF102 JFETs and Q3 is a 2N2222A (see text for additional information about this circuit).

Fig 2-15 shows a recipe type of VFO that you can adapt to your favorite operating frequency upon making a few circuit changes. C1, C2, C3 C4, C5, C7 and C8 are chosen for the operating frequency. L1 and L2 must be changed also. Ballpark values for C4 and C5 are 1500 pF for 160 meters, 1000 pF for 75/80 meters, 560 pF for 30/40 meters and 680 pF for 5 MHz. C3 should be have the least capacitance possible, consistent with reliable oscillator performance. The smaller the C3 value the greater the tuned-circuit isolation from the JFET and its related components. This ensures maximum loaded Q and minimum frequency drift. A great many stable oscillators operate on this principle. The capacitance of C3 is dependent upon the tuned-circuit Q and the transconductance of Q1. The greater the device transconductance and the higher the circuit Q the smaller the practical C3 value.

Q2 serves as an untuned buffer to help isolate Q1 from Q3 and its load. JFET Q2 provides a 100K-ohm load for Q1, which minimizes loading of the oscillator. C6 should be small in value to aid isolation. Good values to use are 100 pF for 160 meters, 68 pF for 75/80 meters, 50 pF for 40 meters and 68 pF for 5 MHz. Smaller C6 values improve the isolation, but at the cost of reduced drive to Q2. Some designers use a resistor (1K to 10K) in series with C6 to improve the stage isolation when using a bipolar transistor at Q2. The disadvantage of this practice is increased circuit noise, which appears in the VFO-chain output. Current that flows through resistance generates noise.

The approximate inductance for L1 can be determined after you calculate the total effective capacitance for the Q1 input circuit. Eq. 2-1 shows how this may be done.

$$Eq. 2-1 \quad C(\text{total}) = \frac{1}{\left[\frac{1}{C_3} + \frac{1}{C_4} + \frac{1}{C_5} \right]} + C_1 + C_2 + 6$$

where C is in pF and 6 is the approximate input capacitance of a JFET. This is with C1, the main-tuning capacitor, fully meshed for operation at the lowest VFO operating frequency. Once the total effective capacitance is known you can determine the required L1 inductance from

$$Eq. 2-2 \quad L(\mu\text{H}) = \frac{10^6}{[2\pi f]^2 \times C(\text{pF})}$$

An NPO ceramic or muRata NPO plastic trimmer may be placed in parallel with C1 to aid in VFO range calibration. Its midrange capacitance should then be included in Eq. 2-1. Alternatively, you may use a slug-tuned coil for L1. You may learn the upper frequency limit of your VFO by applying Eq. 2-1 and using the minimum capacitance of C1 in the formula.

C1 should be a free-turning, double-bearing variable capacitor. This will help to prevent frequency hopping and drift. Avoid using variable capacitors that have aluminum plates for C1. Aluminum expands and contracts with changes in temperature, and this enhances drift. Plated brass vanes are more stable.

The output low-pass network at Q3 in Fig 2-15 is also chosen for the VFO operating frequency. It is designed to match the 800 ohms to 50 ohms, and is not a particularly critical pi network. I use a design Q of 4. This results in an output bandwidth that is adequate for up to 500 kHz of frequency range. The network is designed to cut off slightly above the highest VFO operating frequency, such as 4.5 MHz for a 75-meter VFO. The X_C and X_L for C7, C8 and L2 is 200 ohms. You can calculate the capacitance by using $C(\mu F) = 6.28 \times f_{c0} \times X_C$, when f_{c0} is the cutoff frequency in MHz and X_C is the capacitive reactance of C7 and C8 in ohms. Thus, for a cutoff frequency of 7.5 MHz we find that C7 and C8 should be 100 pF.

The inductance for L2 is obtained from $L(\mu H) = X_L / 6.28 \times f_{c0}$, where 6.28 is 2 times pi and f_{c0} is the cutoff frequency in MHz. Hence, for a 7.5-MHz cutoff we will use 4.2 uH of inductance at L2. X_L is the inductive reactance of L2 in ohms (200 in this example).

D1 in Fig 2-15, as discussed earlier in this chapter, functions as a bias stabilizer for Q1. It limits the transconductance of Q1 on positive peaks of the RF sine wave and helps to minimize harmonic output from Q1. A 1N914 small-signal diode is suitable.

JFET Selection for a VFO

High-transconductance JFETs or MOSFETs are best suited to VFO operation. Of major importance is the "pinch-off" characteristic of the FET. This is commonly referred to as V_p . In common language, pinch off is the value of the drain-source voltage at which no further drain current flows. $V_{gs(off)}$ may be taken to mean essentially the same thing as V_p , except that this condition is of opposite phase to V_p . For our applications of JFETs in oscillators this means that the higher the pinch-off voltage the greater the oscillator output power for a specified drain supply voltage. Current-limiting within the FET does not occur as quickly (during the sine-wave period) as when we use FETs with a low pinch-off characteristic. The 2N4416 JFET is much better than the MPF102 in this respect.

The FET we use in an oscillator should be designed for a maximum frequency that is substantially higher than the proposed operating frequency. The 2N4416 is suitable for use as high as the UHF spectrum. It is, therefore, a good choice for MF and HF oscillator service. The 3N211 and the RCA 40673 dual-gate MOSFETs are both excellent devices for use in VFOs. They may be used as JFETs by tying gates 1 and 2 together to form a single gate, or they can be used with the signal on gate no. 1 and forward bias on gate no. 2.

Short- and Long-Term VFO Drift

Our concern is mainly for long-term frequency drift. Short-term drift occurs during the first two or three minutes after operating voltage is applied. It is caused by the heating of the transistor junction (which causes an internal change in capacitance and resistance). Long-term drift, on the other hand, may last for hours in a poorly designed VFO. A good VFO should settle down within 15 to 30 minutes after being turned on. Long-term drift is caused by RF currents that flow through the VFO capacitors, coil and resistors. These currents cause internal heating of the components and subsequent changes in value. Changes in air temperature around the frequency-sensitive components of a VFO contribute

to long-term drift. A good VFO should not have more than, say, 300 Hz of long-term drift. Short-term drift should not exceed 1 kHz. I have seen some poorly designed VFOs that exhibited up to 10 kHz of short-term drift. These same VFOs had a substantial amount of long-term drift.

The interior area of fixed-value VFO capacitors should be fairly large in order to minimize the effects of RF-current heating. Please refer to Fig 2-15. I have learned that using capacitors in parallel at C3, C4 and C5 provides greater interior surface for the capacitors, and this reduces drift. In a like manner you can use two or three capacitors in parallel at C2 to reduce the effects of internal heating. This recommendation is based on the use of NPO ceramic or polystyrene capacitors.

The Fig 2-15 VFO may be operated also as a series-tuned Colpitts type. C1 and C2 would be placed in parallel between the lower end of L1 and ground. The series configuration is often a better choice at 7 MHz and higher, because the inductance of L1 becomes quite small with all of the shunt capacitance that is present in the circuit of Fig 2-15. Series tuning requires considerably more coil inductance than when using parallel tuning. If the L1 inductance is small, the circuit-board conductors tend to become a significant part of the coil. This causes drift and lowers the coil Q. Eq. 2-1 is not suitable for calculating the total effective C in a series-tuned circuit. It must be modified to include C1 and C2 as series components.

TRANSISTOR MIXERS

Single BJTs or FETs work okay as mixers for nonstringent applications, but they are poor performers in comparison to balanced mixers that use two or four devices. Single-ended mixers have poor dynamic range and offer negligible suppression of the local-oscillator energy. The mixer output usually contains a high level of LO (local oscillator) component. These simple mixers are, however, useful for inclusion in simple bare-bones receivers and similar circuits. They provide an advantage over diode mixers because they have conversion gain rather than loss. This minimizes the number of receiver stages that are needed to provide the necessary 80 dB or greater of overall receiver gain. Examples of simple mixers are given in Fig 2-16.

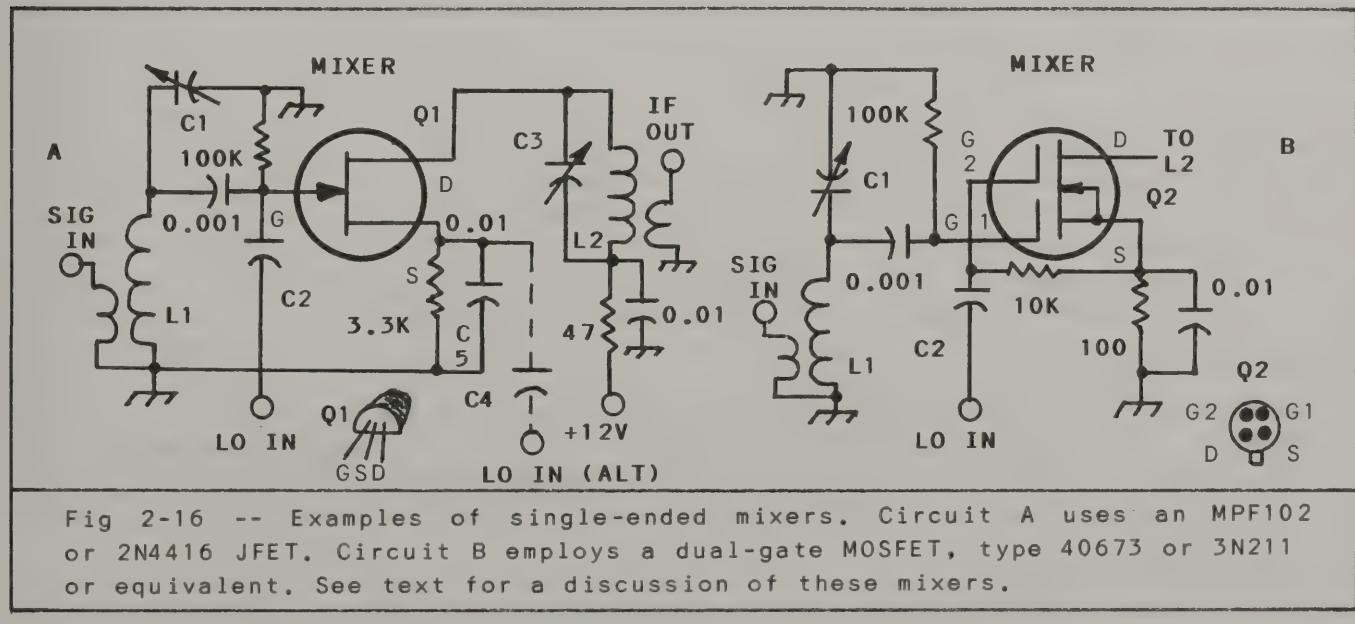


Fig 2-16A illustrates a simple JFET mixer that has the LO (local oscillator) injection applied to its gate. This method is satisfactory if the input signal and the LO frequencies are separated by two or more MHz. When these frequencies are close to the same frequency, load changes at the mixer input, plus the adjustment of C1, will pull the LO frequency. A better technique is to supply the LO energy to the Q1 source (C4 in dashed lines). The source bypass capacitor is eliminated (C5) when this is done. LO voltage with either method of injection should be 5 to 6 volts P-P. Lower injection levels result in reduced mixer conversion gain and degraded dynamic range. Mixer A is capable of providing up to 15 dB of conversion gain. Input circuit C1/L1 needs to have high Q in order to aid the mixer selectivity (ability to separate signals). C2 or C4 are selected to provide the desired LO injection level. The values for these capacitors is dependent upon the output power of the LO chain. Values as great as 0.001 μ F and as low as 10 pF represent the capacitance range you may find necessary. C1/L1 are tuned to the signal frequency and C3/L2 are tuned to the desired IF (intermediate frequency), which may be the sum or difference frequency of the LO and the input signal.

Fig 2-16B shows how you may configure a dual-gate MOSFET as a single-ended mixer. LO injection is applied to gate no. 2 as shown. Again, 5 to 6 volts P-P is the value to use for best mixer performance. Do not exceed 6 volts P-P when using a MOSFET, since this can puncture the gate insulation of the FET. If this occurs, the MOSFET develops a short circuit and the device functions somewhat as a JFET. Bias for gate no. 2 is obtained from the voltage developed across the Q2 source resistor. Forward bias may be applied to this gate by eliminating the 10K-ohm gate-source resistor and using a 3:1 resistive voltage divider that is connected to the +12-V supply line. This requires a 100K-ohm resistor from gate 2 to ground, and a 330K-ohm resistor from gate 2 to the +12-V line. This mixer can yield up to 15 dB of conversion gain. The dynamic range of both mixers in Fig 2-16 seldom exceeds 70-80 dB, which is entirely adequate for most portable and emergency receivers. Isolation between the input signal and LO energy, with relationship to the IF output signal is very poor compared to that of a balanced mixer.

Although bipolar transistors may also be used as simple mixers, I don't recommend them. This is because they have very poor dynamic range, which is usually on the order of 50-60 dB. On the upscale side of things, however, they often provide greater conversion gain that we can obtain with a FET mixer.

The mixers in Fig 2-16 may be used with a broadband transformer at the output, rather than a tuned circuit. When this is done there is a loss in conversion gain, and the input-output signal isolation worsens. In a like manner, the mixer input circuit may contain a broadband transformer rather than a tuned circuit. If you elect to use the broadband method I strongly recommend that a bandpass filter be used ahead of the input transformer. Another bandpass filter should be used at the mixer output to ensure the necessary selectivity. Filter design tables are published in **The ARRL Handbook**.

Self-oscillation is seldom a problem with active mixers, but it can occur if the input signal and IF are closely related in terms of frequency. The problem is then complicated by the use of source injection, because the source of the FET is no longer bypassed. This means that all three FET terminals are "hot" with RF energy.

Balanced Transistor Mixers

We have already discussed the shortcomings of single-ended mixers and the virtues of balanced mixers. Now, let's discuss balanced mixers in greater detail. We will center our discussion around singly balanced mixers (two transistors that are essentially connected for push-pull operation). A doubly balanced mixer would contain four transistors. The practicality of using discrete devices in so elaborate a mixer is overshadowed by the advantages of using an IC that is designed for doubly balanced mixer service. This is because the use of four transistors requires (for best performance) that each transistor exhibit nearly identical electrical characteristics (matched transistors) and the circuit layout becomes rather critical, since good symmetry is essential to balanced operation. We will discuss IC mixers later in the chapter. Meanwhile, please refer to Fig 2-17.

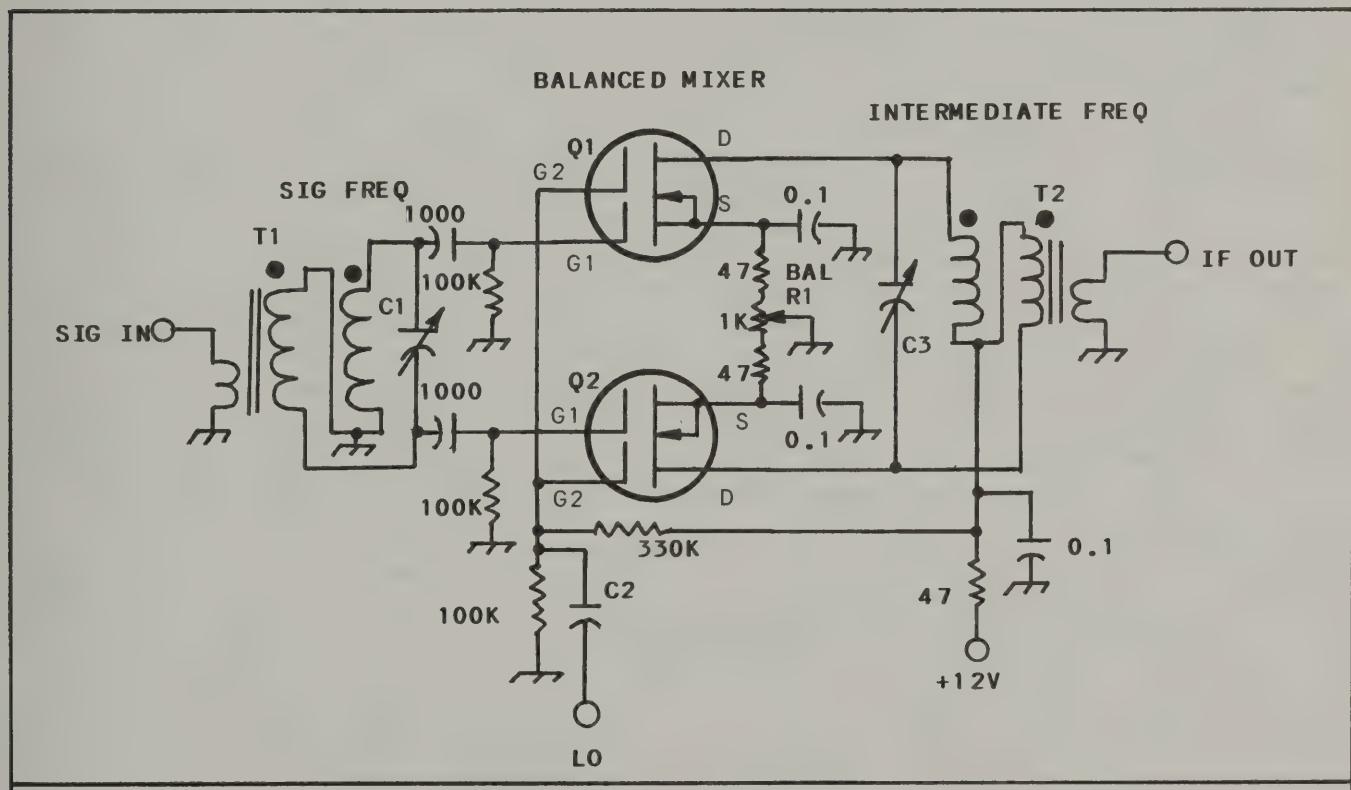


Fig 2-17 -- Practical circuit for a singly balanced MOSFET mixer. C1 and C3 tune T1 and T2, respectively, to resonance at the desired frequency. C2 is chosen, as in Fig 2-16, to provide 5-6 volts P-P at gate no. 2 of each FET. R1 is a trimmer control that is adjusted for best mixer balance, which helps to compensate for transistors that aren't matched. T1 and T2 are bifilar-wound RF transformers. The black dots above the windings indicate the phasing or polarity of the windings. Q1 and Q2 are 40673, 3N211 or 3N212 dual-gate MOSFETs.

The singly balanced mixer yields 10-15 dB of conversion gain with the prescribed value of LO injection. Balance control R1 may be eliminated if the FETs are closely matched. In this event you may tie the sources together and return them to ground through a 270-ohm resistor. A single 0.1-uF capacitor serves as the source-bypass element.

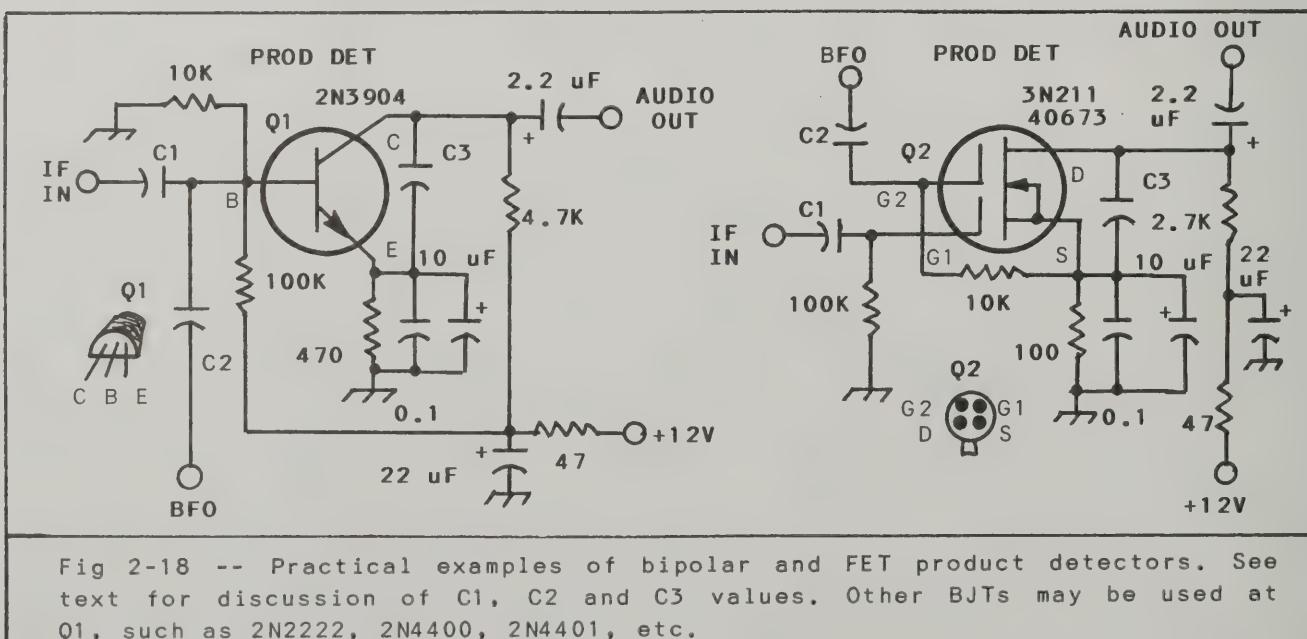
Broadband input and output transformers may be substituted at T1 and T2 if band-pass filters are used ahead of and after the mixer. In either situation, T1 and T2 have a bifilar winding (both main-winding wires placed on the toroid

core, side by side, at the same time). The T1 and T2 link windings are added after the main winding is in place. Powdered-iron toroids are used when the transformers are tuned, as shown in Fig 2-17. The core material should be no. 2 (red) for 1.8 through 7 MHz. Use a no. 6 core (yellow) for 10 MHz through 30 MHz. These cores ensure a high tuned-circuit Q. For broadband mixers I suggest ferrite toroids with a permeability of 800-900. Amidon Assoc. no. 43 core mix is suitable.

JFETs, such as the 2N4416, may be substituted at Q1 and Q2. If this is done you will need to lift the grounded windings of T1 and insert a 100K-ohm resistor between these winding leads and ground. The two 100K-ohm gate resistors and the two gate-blocking capacitors are eliminated. The transformer windings then connect directly to the no. 1 gates. Injection is applied across the 100K-ohm resistor added between the ends of the bifilar winding and ground. No other changes are necessary. The link winding remains unchanged. The Fig 2-17 mixer is capable of a dynamic range of 90 dB.

TRANSISTOR PRODUCT DETECTOR

Product detectors are used for reception of CW, SSB, SSTV and RTTY signals. These detectors are, in effect, mixers. The principal difference between mixers and product detectors is that audio, rather than RF energy, appears at the detector output port. In other words, the output is the product of the two input signals. For example, a receiver that has a 9.000 MHz IF and BFO that operates at 9.0001 MHz will create an audio tone of 1000 Hz (1 kHz) during CW reception. A mixer, conversely, produces an RF (called the IF) output signal that is the sum or difference of the two input signals. Fig 2-18 shows two transistor product detectors. These circuits are used commonly as the input stage of DC (direct-conversion) or synchrodyne receivers, although better performance will be experienced (AM signal rejection) when balanced product detectors are used in DC receivers.



The product detectors in Fig 2-18 produce 10-15 dB of conversion gain with proper BFO (beat-frequency oscillator) injection. For circuit A this should be on the order of 3-4 volts P-P and for circuit B the desired level is 5-6 volts P-P. C2 in each circuit is selected to ensure the desired BFO voltage. The value is dependent upon the output power of the BFO. Values from 10 to 1000 pF are typical. Measurements may be made with a scope or RF probe. P-P voltage may be converted to RMS (root mean square) by multiplying the P-P voltage by 0.3535. An RF probe reads RMS voltage, hence the need for conversion.

C1 in Fig 2-18 is also chosen for the prevailing circuit conditions. If the product detector is connected to a low-impedance point at the output of the last IF amplifier (such as to the link of an IF transformer) we may use fairly heavy coupling. Q1 has an input impedance of roughly 1000 ohms, as shown. Therefore, in a low-impedance circuit we may use C1 values from 0.001 to 0.1 uF. If, on the other hand, we connect the product detector to a high-impedance point at the IF amplifier output (such as the top of the main winding of the IF transformer), we must use light coupling. C1 values from 10 to 220 pF are typical. The lower the IF the greater the capacitance used. You will need to do this experimentally if you aren't a design engineer. In any event, C1 should not be so large that it loads the IF amplifier and causes broad tuning and reduced IF output power. Q2 has an input impedance of 100K ohms.

C3 in Fig 2-18 has no assigned value. Here again we may make an arbitrary choice for the capacitor value. C3 serves two purposes: (1) it bypasses the BFO energy that appears on the collector or drain of the product detector, and (2) it bypasses the high-frequency audio energy. The receiver audio response may be shaped in part by choosing a particular value of C3 capacitance. It is beneficial to roll off the highs at 1000 Hz for CW reception and 2500 Hz for SSB reception. This eliminates receiver hiss noise and greatly attenuates high-pitched QRM. Typical C3 values range from 0.005 to 0.1 uF.

You will observe that the emitter of Q1 (fig 2-18A) and the source of Q2 (Fig 2-18B) are doubly bypassed. The 0.1 uF is for RF energy and the 10-uF capacitor is for audio frequencies. The 10-uF capacitor ensures maximum audio output from the product detectors.

TRANSISTOR DC AMPLIFIERS and SWITCHES

Transistors are used frequently as substitutes for mechanical switches. These electronic switches help to eliminate troublesome switching transients (voltage spikes caused by the breaking of contacts when current is flowing). They also eliminate the need for long leads in signal circuits, such as those from the contacts of a mechanical switch to and from a circuit board. Long switch leads can cause circuit instability and degraded tuned-circuit Q. Electronic switches may be used for audio muting, keying circuits and a host of other applications. Diodes are more commonly used for switching coils or filters, but BJTs may be employed for this job too.

Dc amplifiers are used to increase the sensitivity of a meter, as AGC (automatic gain control) amplifiers and in similar circuits. Most receiver S-meter circuits have a dc amplifier fed from the AGC line or one of the audio amplifiers.

You may use NPN or PNP transistors as dc amplifiers and bipolar switches. The choice is contingent upon the type of circuit to be switched or amplified. Fig 2-19 shows how to use transistors to key a CW transmitter and drive a relay.

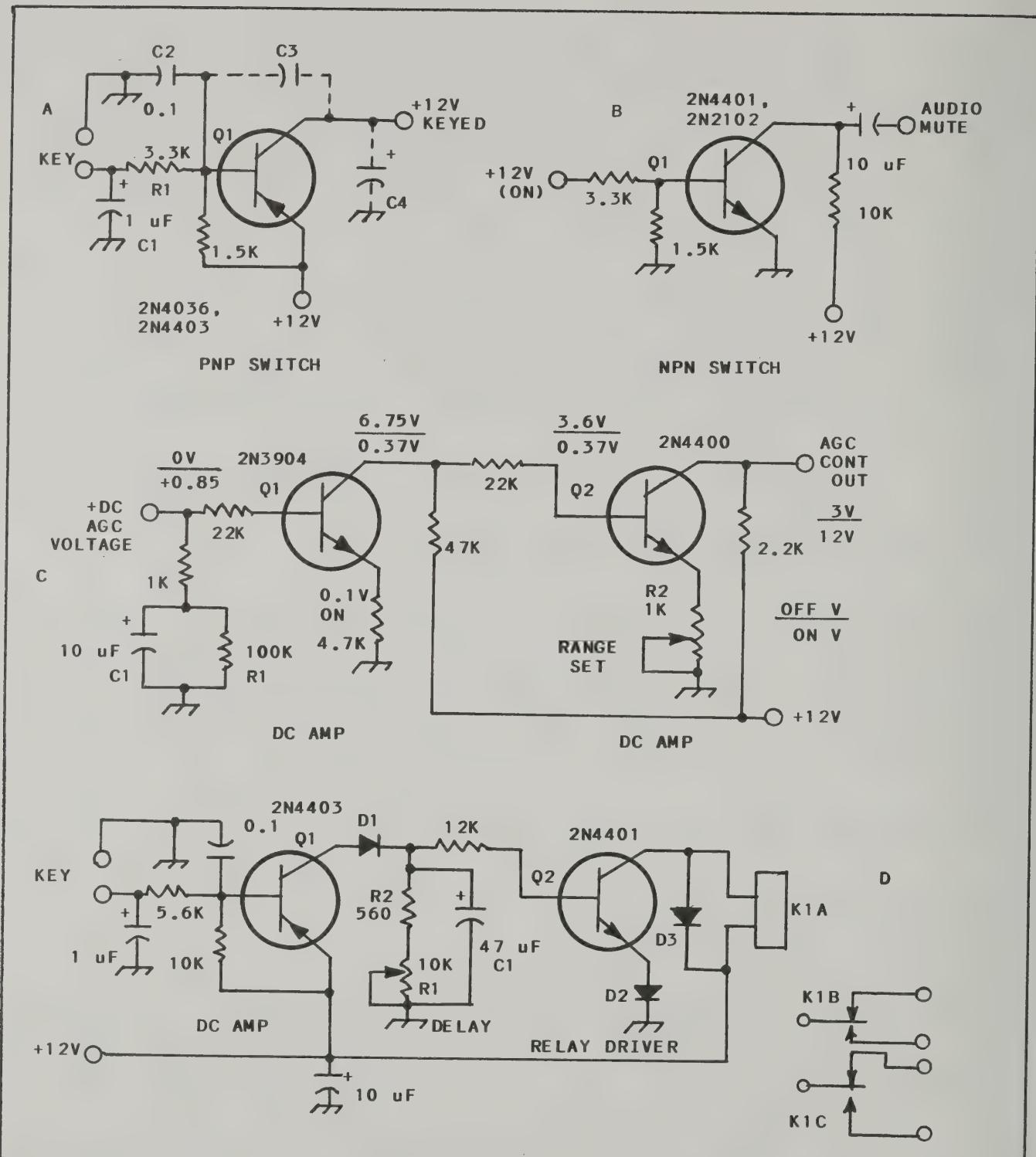


Fig 2-19 -- Circuit A is a PNP switch that is suitable for keying a low-power stage in a CW transmitter. It may be used also for turning on various +dc supply lines. Circuit B shows a PNP switch which is configured to mute an audio line in a receiver. An AGC control circuit is shown at C. It is designed to work with MC1350P or MC1590 IF amplifier ICs. A break-in delay circuit is illustrated at D. It may be used in a CW transmitter to switch the antenna and mute the receiver by using the contacts on K1, which is a 12-V dc relay. Consult the text for a more complete explanation of the above circuits.

The circuit of Fig 2-19A is suitable for turning on and off any dc supply line by means of a key or switch. When it is used for keying a low-level stage of a CW transmitter you may shape the keyed RF output waveform by choosing the proper component values for C1, C2, C3, C4 and R1. This is beneficial when you want to eliminate key clicks. The shaping components affect the rise and fall times of the keyed waveform, which should be approximately 5 milliseconds for a quality wave shape. The C1 and C2 values are generally as shown in the circuit. C4 can be any value from 1 μ F to 10 μ F that provides the desired shaping. C3 is generally much smaller in value (0.01 to 0.1 μ F), consistent with the shaping you desire. If R1 has too high a resistance you will not be able to cause Q1 to turn on completely. The larger the R1 value the slower the rise time of the keyed waveform. Values from 2.2K to 10K ohms are typical. Switching transistor Q1 is actuated by grounding the junction of C1 and R1. The transistor must be capable of carrying the current of the circuit to be keyed without its ratings being exceeded. It should also have a fairly high dc beta (100 or higher).

An NPN switch is illustrated at B of Fig 2-19. Turn-on is accomplished by applying forward bias to the base of Q1. This switch may be used for shunt bypassing of an audio line for receiver muting. When Q1 conducts it allows the 10- μ F output capacitor to return to circuit ground through the Q1 collector-emitter junction. When Q1 is in the off state, the capacitor is effectively out of the circuit.

A transistor AGC-control system is presented at C of Fig 2-19. It requires a dc control voltage. It starts to function when 0.85 volt reaches the base of Q1. The control voltage is obtained by rectifying signal energy from the last IF in a receiver, or from one of the low level audio-amplifier stages. This system may be used for audio- or RF-derived AGC, depending upon your preference. An AGC amplifier should be used between the signal sampling point and the AGC control circuit shown. The IF or audio energy is rectified by a 1N914 or similar diode before it is applied to Q1. Output from Q2 ranges from +3 to +12 volts. This range is proper for a Motorola MC1350P or MC1590 IF-amplifier IC, for which +3 volts produces maximum IC gain. C1 and R1 determine the AGC time constant. The values shown yield a 1 second decay time. R2 is adjusted for the desired Q2 output-voltage range. Voltages are shown at key points in the circuit. The upper notations are for operating voltage when there is no rectified signal energy applied to the base of Q1. The 22K-ohm resistors in series with the dc to the bases of Q1 and Q2 tend to linearize the AGC control-circuit response. An S-meter circuit may be operated from the control voltage at the output of Q2.

Fig 2-19 illustrates a practical break-in delay system for use with a CW transmitter. Q1 is keyed on by grounding the input circuit. A charge is placed in C1 and it discharges in accordance with the combined resistance value for R1 and R2. R1 is adjusted for the desired drop-out time. Upon closure of the key, Q2 is saturated and current flows through relay K1, causing closure of the relay contacts. K1B and K1C may be used for antenna switching and receiver muting. D1 prevents damage to Q1 during key-up (base open) when C1 is fully charged. D2 establishes 0.7 volt of bias to ensure quick release of K1. D3 clips transients caused by the collapse of the K1 field coil when the key is opened. A 12-V DPDT DIP relay is excellent for use at K1. This circuit operates in a manner similar to that of a VOX circuit, and is adaptable to that application by applying rectified audio to a transistor switch that shorts the Q1 input network to ground when the operator speaks into the mic. You can observe that the circuit requires a PNP and an NPN switch. Q2 must be able to safely accommodate the current that is drawn by K1.

VOX CIRCUIT

The break-in delay circuit shown at D of Fig 2-19 may be changed to function as a VOX (voice-operated relay) module if you anticipate building an SSB transmitter. All that is required to convert the circuit to a simple VOX unit is to add an audio amplifier and rectifier ahead of the existing circuit. The audio energy from the mic is rectified and the resulting dc voltage actuates an NPN switch that keys Q1 in Fig 2-19D. A suggested circuit is shown in Fig 2-20.

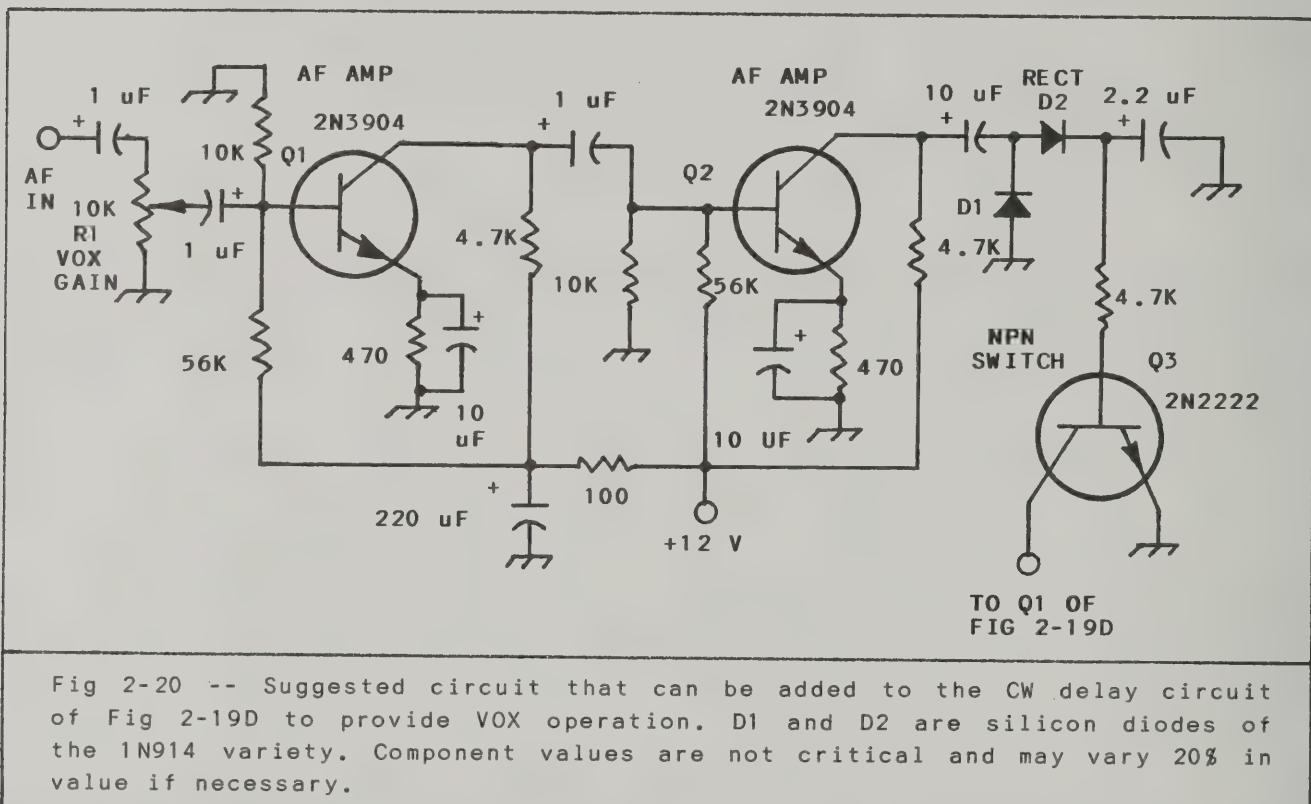


Fig 2-20 -- Suggested circuit that can be added to the CW delay circuit of Fig 2-19D to provide VOX operation. D1 and D2 are silicon diodes of the 1N914 variety. Component values are not critical and may vary 20% in value if necessary.

Input for this circuit is obtained from the mic audio line or from the output of one of the speech-amplifier stages in the SSB transmitter. R1 is adjusted to ensure VOX actuation at normal speech levels and operator distance from the mic. Q1 and Q2 amplify the voice energy, which is rectified by the D1, D2 doubler circuit. The resultant dc voltage is routed to the base of NPN switch Q3. The plus voltage causes Q3 to conduct, and this causes the collector-emitter junction to present a low resistance to Q1 of Fig 2-19D, thereby turning on the relay, K1, in Fig 2-19D. VOX drop-out time is set by means of R1 in Fig 2-19D.

You may wish to bypass the bases of Q1 and Q2 in Fig 2-20 with 0.01-uF capacitors. This will aid in keeping unwanted RF energy out of the VOX circuit. Stray RF energy can cause erratic VOX operation. A miniature 500-uH or 1-mH RF choke can be placed in series with the audio line to the Q1 base if RF problems persist. This choke should be located as close to Q1 as practicable.

CHAPTER SUMMARY

The circuits in this section serve as an introduction to the use of discrete active devices. Simplicity is the keynote, and component values are not generally critical. Likewise with your choice of transistors. A transistor substitution guide is helpful when selecting alternative transistors.

Glossary of Chapter Terms

AV - Amplifier voltage gain, frequently expressed in dB.

Balanced (mixer) - A passive or active circuit that has two or four semiconductor devices (diodes, transistors or an IC) for use as a mixer in a receiver or transmitter. Two devices in a push-pull type of arrangement comprise a singly balanced mixer, whereas four such devices form a doubly balanced mixer. A single-ended mixer has but one device.

Bipolar Switch - A transistor that is used in place of a mechanical switch to turn on or off a circuit. Forward bias is applied to the transistor when it is turned on.

Boost circuit - Used to accentuate the high- or low-frequency audio energy in a microphone amplifier or Hi-Fi amplifier, etc.

Break-In Delay - A circuit that turns on a CW transmitter when the key is closed. A capacitor in the delay circuit is charged to cause a slow release (drop out) of the control circuit, which is usually a relay. A potentiometer in parallel with the charged capacitor sets the drop-out time of the circuit.

Broadband amplifier - An RF amplifier that has uniform gain over a broad frequency spectrum, such as 1.8 to 30 MHz. It requires no tuning when the operating frequency is changed.

Buffer amplifier - A transistor stage that follows an oscillator. It helps to isolate the oscillator from the subsequent stages in a transmitter or receiver. A buffer may or may not provide gain, depending on the design and circuit requirements.

Combiner - A broadband transformer with two input and one output terminal or port. Used for feeding the output of two RF power amplifiers into a single load, such as a filter or antenna.

Conversion gain - The gain of a mixer, expressed in dB, over the level of the input signal applied to the mixer. If the mixer gain is less than the value of input signal, the mixer has a conversion loss, such as with diode mixers.

DC amplifier - A transistor or IC amplifier that has dc applied to its input terminal. It provides a current gain rather than a signal gain.

Discrete device - A single diode or transistor. An IC is not a discrete device, since it contains many diodes and transistors.

Drift, oscillator - An unwanted slow or rapid change in oscillator operating frequency. Drift results from changes in component temperature, caused by variations in ambient temperature and RF circulating current.

Efficiency, amplifier - The amplifier input power versus the useful output power as expressed in percentage.

Electret, microphone - A microphone element that requires an operating voltage. Also called a "condenser mic."

Feedback-- Amplifier output energy that is fed back to its input circuit. Causes an amplifier to oscillate and is required for a circuit that is designed to function as an oscillator.

Heat sink - A metal device to which a transistor or IC is attached for the purpose of cooling the internal workings of the device. Also used with power diodes for the same purpose. Usually made from extruded aluminum.

Hi-Z - Used to signify high impedance of an arbitrary value.

LC circuit - A resonant circuit that contains capacitance (C) and inductance (L) in parallel or series. Generally referred to in connection with oscillators and filters.

Linear operation - Generally associated with amplifier operation. If the amplifier is linear it reproduces faithfully at its output the same wave form that is applied to its input terminal. This ensures minimum distortion at the amplifier output.

LO - Refers to local oscillator. LO injection is oscillator output voltage that is applied to a mixer, for example.

LO-Z - Signifies low impedance of an arbitrary value.

Mixer - An active or passive device. Combines two input energies to produce a third one that is the mathematical sum or difference of the two input frequencies.

MMIC - Monolithic microwave integrated circuit. Designed specially for use in the upper UHF and microwave spectrum. Features low noise and high gain.

Narrow band, amplifier - The opposite of a broadband amplifier. An amplifier that contains one or more tuned (resonant) circuits, which restrict the frequency response to a narrow range in kHz.

Overtone, oscillator - A crystal oscillator that causes the quartz crystal to oscillate at odd harmonics of the crystal fundamental frequency, such as the 3rd, 5th and 7th overtones.

Parasitic, oscillation - An unwanted amplifier self-oscillation that may occur at frequencies well removed from the amplifier design frequency.

Product detector - Used in receivers to demodulate SSB signals. The output is the product of the two input signals (IF and BFO signals) and occurs at audio frequency. Operates in a manner similar to that of a mixer.

Splitter - A broadband transformer that has one input and two output ports or terminals. Used to equally divide an input signal into two identical outputs.

Switch, transistor - Used in place of a mechanical device to turn on or off another circuit. An IC or a single transistor may be used as a switch.

VOX - Voice operated relay in a transmitter.

VXO - A crystal oscillator containing circuitry that permits the crystal frequency to be shifted above and below its marked frequency.

DIODE & IC APPLICATIONS

CHAPTER 3

I wish to continue in this chapter with our recipe-book approach to circuits that are useful to the experimenter and radio amateur. In this section we will discuss diodes and ICs in terms that you can understand. I want to include in this chapter a collection of practical circuits with assigned component values. We shall emphasize analog applications in our circuits, since if I were to cover digital circuitry as well, we would run out of page space quickly! You will find a comprehensive treatment of digital and logic ICs and related theory in **The ARRL Electronics Data Book**, second edition, 1988.

DIODES and ICs in GENERAL

The earliest form of radio detector was a diode. The first radios for reception of commercial AM broadcasts contained nothing more than a diode detector, a tuned circuit and a pair of headphones. Diodes continue to be used as detectors in many modern receivers, but their structure and general performance have taken a forward leap since the days of galena crystals and catswhiskers! We now have a wide assortment of silicon and germanium diodes to choose from when designing a circuit. We also have the hot-carrier or Schottky diode to add to the list. The list would not be complete without Zener or reference diodes, along with varactor diodes. Carborundum, copper-oxide and selenium diodes have been buried in the dust of past decades. Most vacuum tubes diodes, such as the 6AL5, 6H6 and 5Y3 have been replaced by solid-state equivalents.

Integrated circuits, in general, have replaced entire complex circuits rather than one discrete device. ICs contain literally hundreds of transistors, diodes, resistors and capacitors. These components are formed on a common piece of silicon crystal that is called the substrate. A single IC may occupy one square inch of space on a circuit board, whereas an equivalent circuit with discrete transistors and the related components might require a square foot of circuit-board area. LSI chips (large scale integration) may be, for example, 5 inches long and an inch wide. The major electronic workings of a computer can be contained in a single LSI device. I can recall helping to design the MIDAC computer in the early 1950s, which used vacuum tubes. The composite computer was housed in 6-foot high cabinets that occupied some 15 feet of wall space! Times have changed considerably!

We have linear and logic ICs today. We amateurs use linear ICs for most of our work in receivers and transmitters. Logic ICs are used in frequency counters, digital frequency displays, voice synthesizers, and such. Both families of ICs play a vital role in amateur radio circuits.

Some ICs are fragile and can't be subjected to static electricity without being destroyed. These ICs contain MOS (metal oxide silicon) devices that are unable to handle small static-voltage spikes without the thin layer of insulation becoming

punctured. Therefore, we must handle the ICs with extreme care. It is wise to wear a grounding strap on our wrists while handling these ICs. In a like manner, the pencil soldering iron tip should be connected to earth ground when we solder MOS ICs into a circuit board.

Although we have the option of using IC sockets on our PC boards, it is better to hard wire (solder directly) the ICs into the circuit board. Sockets tend to develop resistive contacts from exposure to air pollutants, and this leads to intermittent operation or complete circuit malfunction. Sockets offer the advantage of easy IC replacement and they are often preferred for experimental breadboard types of circuits. Low-profile (low height) IC sockets are the best for audio and RF circuits because they minimize the effective length of the pins on the IC. Long leads can cause low gain and instability.

DIODE DETECTORS, BALANCED MODULATORS, PRODUCT DETECTORS & MIXERS

Diodes offer high performance in terms of dynamic range (strong-signal handling ability) when they are used in mixers, detectors and balanced modulators. They have a conversion loss rather than a gain, which is the "down side" of using them in these circuits. When transistors or ICs are used for the same applications we obtain a conversion gain. This is useful in minimizing the number of gain stages in a circuit: More gain is often necessary to compensate for the loss through a diode mixer. Losses on the order of 5 to 8 dB are common for a balanced diode mixer or balanced modulator. The advantages of increased dynamic range for these passive circuits often outweighs the conversion loss. An active equivalent circuit may produce up to, say, 15 dB of conversion gain. But, active mixers can't accommodate large signals without a degradation in dynamic range.

A good active mixer may not require an RF amplifier stage ahead of it for obtaining a satisfactory receiver noise figure, at least up to 10 MHz. The antenna noise usually exceeds that of the active mixer below 14 MHz. Most diode mixers do require the addition of an RF preamp in order to have the signal override the mixer noise. These considerations are seldom germane to the design of balanced modulators and product detectors, where gain is normally available in abundance external to the balanced modulator or product detector. This is also true of balanced mixers that are used in transmitter circuits.

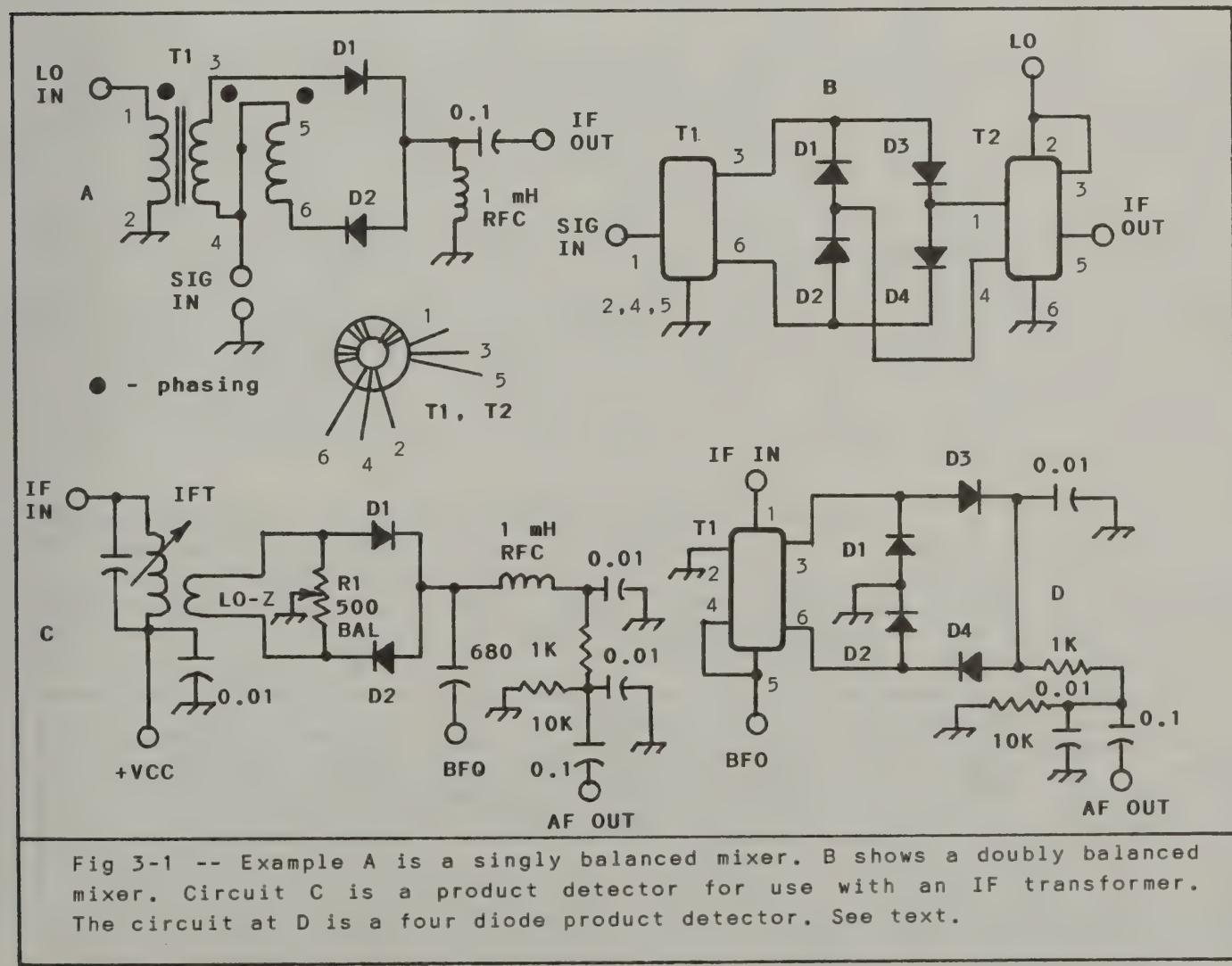
You can build your own balanced mixers, balanced modulators and product detectors if you match the diodes for forward resistance. Small-signal diodes, such as 1N914s, are suitable for most amateur circuits. They are inexpensive and widely available as surplus. Hot-carrier diodes of the HP2800 (Hewlett-Packard brand) are perhaps the best choice for diode balance and general performance, but they are costly compared to the generic 1N914 silicon diode. The Schottky diodes sold by Radio Shack are good performers too.

Diode matching is done by measuring the forward resistance of the diodes with a VOM or VTVM. The typical forward resistance ranges from 5 to 15 ohms. Select a set of diodes that have the same forward resistance, and the closer the resistance the better the circuit balance. The back resistance of the diodes is usually 1 megohm or greater for silicon devices, and 100K ohms or greater for germanium diodes. The back resistances need not be matched, since they do not play an important role in circuit balance. The forward resistance is that which is measured from the diode anode to its cathode.

Broadband transformers are almost always used at the input and output of diode

balanced mixers and modulators. This type of transformer is used also at the input of a diode product detector. The 10-Z secondary of an IF transformer may also be used to feed a diode product detector if a resistive divider is used across the secondary winding to establish an electrical center tap (balance). This is often a 500-ohm potentiometer that is adjusted for circuit balance. The arm of the control is grounded in this application.

Diode mixers, balanced modulators and product detectors have an input and output impedance on the order of 50 ohms. Therefore, these circuits must be matched to the energy source and load in order to minimize mismatch losses. Trifilar broadband transformers are used for input and output coupling to these diode circuits. The three windings are laid on the toroid core at the same time, and they are each the same length. The wires may be placed on the core in parallel (side by side) or they may first be twisted together (6-8 twists per inch) and then wound on the core. If you use a different color enamel wire for each conductor it will be easier to identify the three wires when they are connected to the circuit. I often spray the ends of the wires with paint (different color for each wire) before I place the wire on the core. This makes identification a simple matter. For operation from 1.8 to 50 MHz I use 15 trifilar turns of no. 28 enamel wire on an Amidon Assoc. FT-37-43 ferrite toroid. This core has a 0.37-inch OD and a permeability of 850. Fig 3-1 shows various circuits in which diodes are used with reference to the foregoing discussion.



The simple mixer of Fig 3-1A offers good performance, but does not have the good port-to-port signal isolation that is available from the doubly balanced mixer at B in the same figure. Circuit B is the same as that which is built in modular form by companies such as Mini Circuit Labs. The diodes in these mixers should be balanced carefully, as discussed earlier. They may be 1N914 or HP2800 devices. The broadband trifilar transformers in Fig 3-1 (T1 and T2) contain 15 turns of no. 28 enamel wire on an Amidon Assoc. FT-37-43 ferrite toroid. LO injection power for these mixers should be +7 to +10 dBm (5 to 10 mW) for best mixer performance.

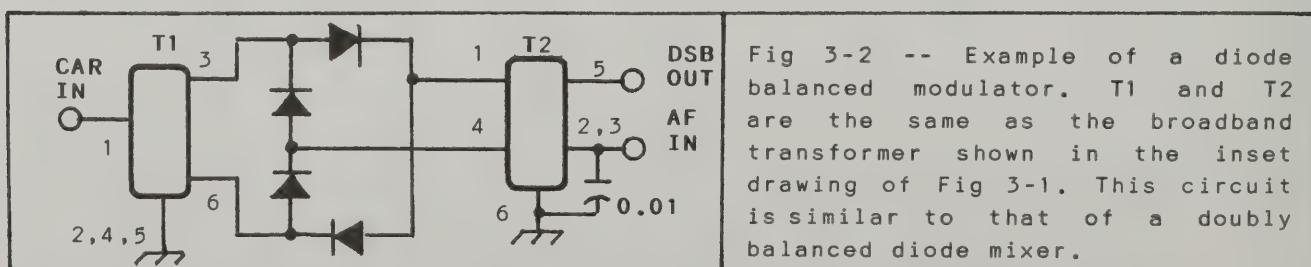
Fig 3-1C illustrates a two-diode product detector that is suitable for use with a standard IF transformer. The diodes connect to the low-impedance transformer secondary, and R1 provides the circuit center tap. This control is adjusted for detector balance. It may be omitted if you replace it with two 270-ohm resistors. The common point for the resistors is grounded. The RF choke, 1K-ohm resistor and the related bypass capacitors prevent BFO and IF energy from reaching the audio-amplifier circuit.

The four-diode product detector at D of Fig 3-1 uses a broadband transformer at the input. The audio output is filtered by means of the 1K-ohm resistor and the two 0.01-uF bypass capacitors. BFO injection for both product detectors is +7 to +10 dBm, as is the case with the two mixers in Fig 3-1. The term "dBm" is referenced to power in mW, with 0 dBm being equal to 1 mW.

Balanced modulators that use diodes are the same as the two mixer circuits in Fig 3-1. The principal difference is that audio voltage is applied to one of the ports, while carrier-generator energy from an internal local oscillator is supplied to the second port. The third (output) port provides suppressed carrier SSB energy that is routed through an SSB crystal or mechanical filter. See Fig 3-2.

There is considerable confusion about the phasing of the broadband transformer windings. The large black dots (see Fig 3-1) indicate the transformer phasing or polarity. The dots are located at the start of each of the windings. In other words, leads 1, 3 and 5 of the inset pictorial drawing are the starts of the three windings. If the indicated phasing is not observed, the circuit will not function in a balanced manner. D1 and D2 at A of Fig 3-1, for example, must be supplied with energy that has a 180 degree phase difference. Note that windings 3-4 and 5-6 are reverse-connected in order to provide the proper phase or sense. This may be thought of as a push-pull arrangement.

In all of the Fig 3-1 circuits it is essential to maintain the best physical and electrical symmetry possible in order to enhance the circuit balance. All leads must be kept short and direct, consistent with symmetry. The mixer circuits, in particular, need to be contained in shielded enclosures to prevent stray signal energy from entering them beyond the normal input port. This is not a vital consideration for product detectors.



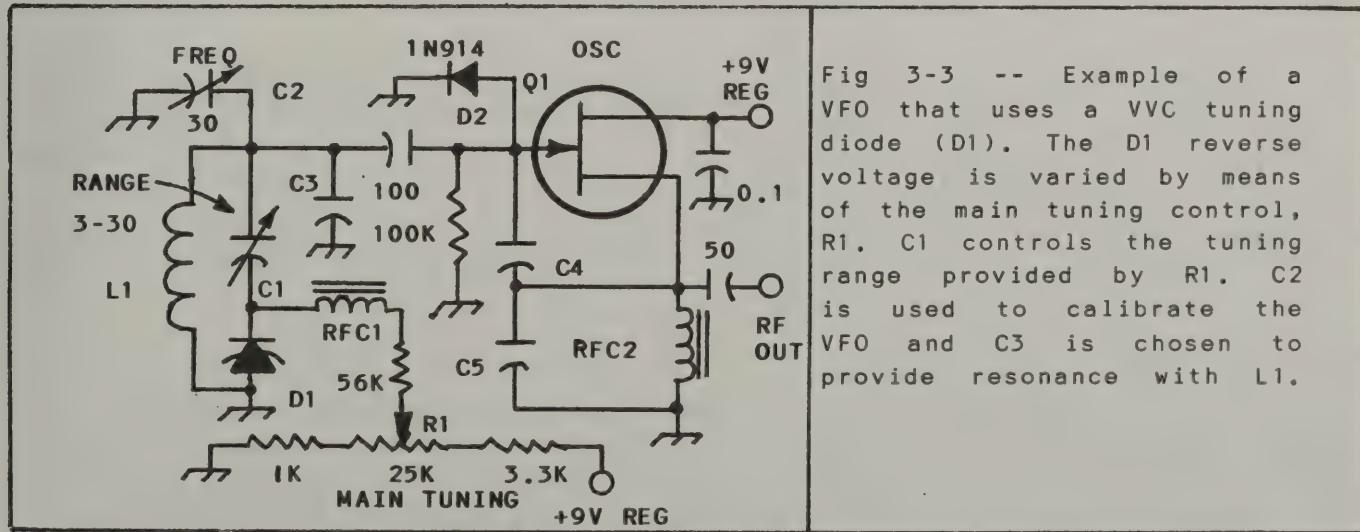
VVC or Tuning Diodes

Certain types of diodes may be used as electronic variable capacitors. These are known as VVC (voltage variable capacitance) or tuning diodes. You will hear them referred to also as varactor diodes. These diodes may be used in place of mechanical tuning capacitors in many circuits. This is a benefit when we wish to reduce the cost of a project, and it permits us to reduce the size of the equipment.

It is necessary to apply reverse voltage to a VVC diode in order to make it work as a capacitor. Specifically, we apply a positive voltage to the diode cathode while the anode is grounded. As the reverse voltage is varied there is a change in the diode junction capacitance. A potentiometer is used to change the applied voltage.

The principal disadvantage of VVC diodes is their sensitivity to internal and external temperature changes. As the temperature changes, so does the effective junction capacitance. This condition causes short-term drift that is often greater than that with a mechanical tuning capacitor. The short-term drift (in an oscillator, for example) generally occurs during the first five minutes of diode operation, during which time the diode junction heats up from the flow of dc current. This phenomenon is not a major problem in amateur circuits if we allow for the warm-up period before calibrating frequency readout dials.

Attention must be paid to the general characteristics of the VVC diode or diodes we select for a circuit. The Q of the diode versus frequency is of importance if we are to prevent the VVC diode from degrading the overall Q of the tuned circuit. The manufacturers' literature lists the Q and recommended upper frequency of use for these diodes. The data sheets also contain a curve that indicates the useful capacitance range of the diode versus applied voltage. Most VVC diodes have a fairly linear capacitance change over a portion of the curve. This is the area of major interest to us. A typical voltage range for a tuning diode is +1 to approximately +15 V dc. The operating voltage should be regulated in order to prevent capacitance changes when the supply voltage varies. The higher the applied voltage the lower the diode capacitance. There is very little change in capacitance below 2 volts. Fig 3-3 shows how we can utilize VVC diodes for oscillator tuning.



D1 in Fig 3-3 is the VVC diode, as indicated by the curved line through the anode. RFC1 can have an inductance between 500 and 1000 μ H (not critical). This choke isolated the dc circuit from the VFO tuned circuit. Please note the limiting resistors at each end of R1. The values are chosen to allow D1 to operate over the most linear portion of its capacitance curve. The values chosen for the two limiting resistors are dependent upon the VVC diode used. A 10-turn Helipot is ideal for use at R1, since it provides vernier tuning of the frequency control. Standard vernier-drive mechanisms may be used also. R1 will yield the smoothest performance and best longevity if you use an industrial grade (such as Allen Bradley brand) 2-W carbon composition control for R1. Low-cost, imported controls become intermittent quickly.

Improved oscillator performance may be had by using two VVC diodes in a back-to-back arrangement. This is shown in Fig 3-4. The oscillator output waveform is more linear when two diodes are used, whereas a single VVC diode (Fig 3-3) clips a portion of the negative half of the RF sine wave. This condition is minimized when we employ two diodes. RF current should not be allowed to flow through the diodes and cause them to conduct. Back-to-back diodes prevent this condition.

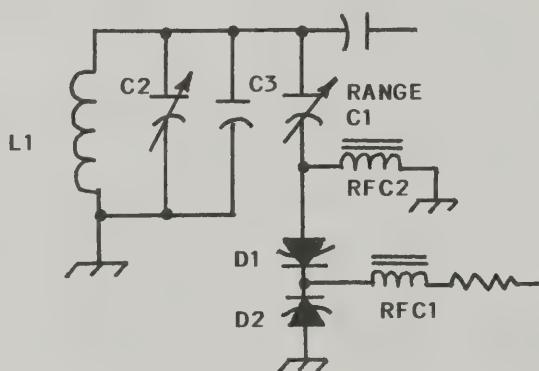


Fig 3-4 -- Circuit example that shows how two VVC diodes may be used to improve the oscillator output wave form. This circuit is similar to the one in Fig 3-3 except for the addition of D2 and RFC2. A dc ground return for D1 is provided by RFC2.

Since D1 and D2 of Fig 3-4 are in series, the net capacitance will be half that of one diode. You may select diodes that have twice the capacitance required for the desired tuning range. The correct amount of effective capacitance will then be available when the diodes are connected as in Fig 3-4.

Small-signal diodes, such as the 1N914, may be used as VVC diodes. Owing to the small junction area of these diodes, the capacitance will be quite low. You may parallel two or more 1N914 diodes to increase the effective capacitance. Bipolar transistors may also be used as VVC diodes. The tuning-control voltage (positive) is applied to the emitter of an NPN transistor and the collector is grounded. Additional capacitance may be had by paralleling two or more transistors. Devices such as the 2N3904, 2N2222A and 2N4400 are good examples of transistors that may be used as VVC diodes.

Diodes as Frequency Multipliers

Varactor diodes are used at VHF and higher as power frequency multipliers. For example, the Motorola MV1805C power varactor is designed for frequency multiplication from 100 MHz to 2.0 GHz. It is rated for 26 watts of output power with an RF input (driving) power of 40 watts. As a tripler from 250 to 750 MHz, for example, it has an efficiency of 65%. No operating voltage is required for a power varactor -- only RF input power. Fig 3-5 shows the circuit for a varactor

frequency doubler. Three or four 1N914 diodes may be connected in parallel and used in place of D1 of Fig 3-5, but the output power will be very low (mW).

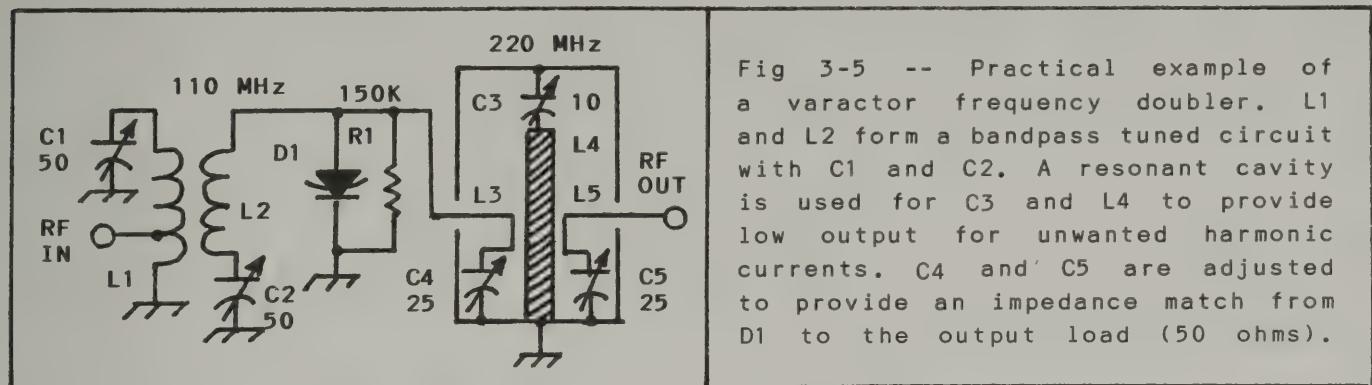


Fig 3-5 -- Practical example of a varactor frequency doubler. L1 and L2 form a bandpass tuned circuit with C1 and C2. A resonant cavity is used for C3 and L4 to provide low output for unwanted harmonic currents. C4 and C5 are adjusted to provide an impedance match from D1 to the output load (50 ohms).

The doubler circuit shown above need not have a resonant cavity at the output, as shown. Lumped inductance may be substituted by using a circuit that is similar to the C1/L1, C2/L2 circuit. The harmonic attenuation will not be as great as when you use a resonant cavity at the doubler output. Each of the capacitors in Fig 3-5 is adjusted for maximum doubler output power with 110-MHz drive applied. D1 may be one of the many varactor diodes produced worldwide. A good example is the MV1805C or MV1809C1 step-recovery diodes. The latter device is rated for operation from 300 MHz to 3.0 GHz and can deliver 17 watts of output power at 2.0 GHz.

Fig 3-6 shows a varactor tripler circuit that is designed for multiplying from 144 MHz to 432 MHz. With 25 watts of driving power the output power is on the order of 15 watts.

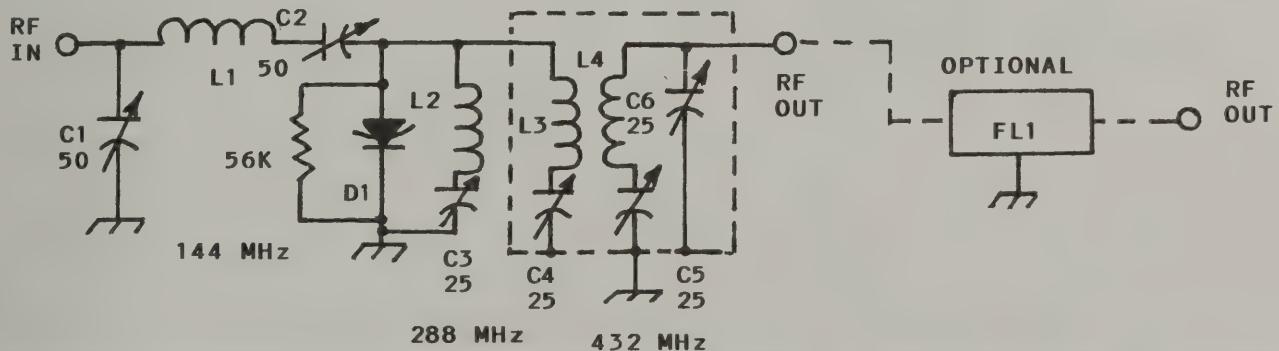


Fig 3-6 -- Circuit for a varactor tripler from 144 to 432 MHz. An optional harmonic filter (FL1) is shown at the right. This can be a resonant cavity of the type depicted in Fig 3-5 if additional harmonic suppression is desired. Approximate coil data is: L1, 5 turns of no. 20, 3/8 inch ID by 1 inch long; L2, 4 turns of no 20, 1/4 inch ID by 1/2 inch long; L3 and L4, 3 turns of no 14, 3/8 inch ID by 3/4 inch long. C4, C5, L3 and L4 are enclosed in a copper or brass shield box. The spacing between L3 and L4 is roughly 3/8 inch.

L3 and L4 in Fig 3-6 should be silver plated for best Q. Likewise if a resonant cavity is used. Cool Amp rub-on silver-plating powder may be used to accomplish this. All of the tuning capacitors in Figs 3-5 and 3-6 are miniature air-variable types. The capacitors in Fig 3-6 are adjusted for maximum output power, consistent with minimum unwanted harmonic output current. A spectrum analyzer is best for this purpose, but you can obtain acceptable results when using a UHF RF wattmeter

and a home-made calibrated VHF/UHF wavemeter or dip meter. The wavemeter is used to sample the doubler or tripler output while the wavemeter is tuned to various harmonic frequencies. There should also be minimum multiplier input-frequency energy present at the multiplier output port. Please note that the 288-MHz tuned circuit in Fig 3-6 is commonly called the "idler tank." Its purpose is to function as a series 288-MHz trap to remove most of the second-harmonic energy of 144 MHz.

You may wish to use a bandpass type of input tuned circuit (Fig 3-5) in the circuit of Fig 3-6. This will help to suppress harmonic energy from the 144-MHz exciter, which in turn will minimize the harmonic current applied to D1. It is important to ensure that all of the tuned circuits in varactor multipliers are of the highest Q practicable. Wire gauges lower than those specified will result in greater values of tuned-circuit Q.

Small-signal diodes may be used successfully as frequency multipliers at HF as well. Overall current can be reduced in a circuit by using diode multipliers. Two examples of low-level frequency multipliers are presented in Fig 3-7.

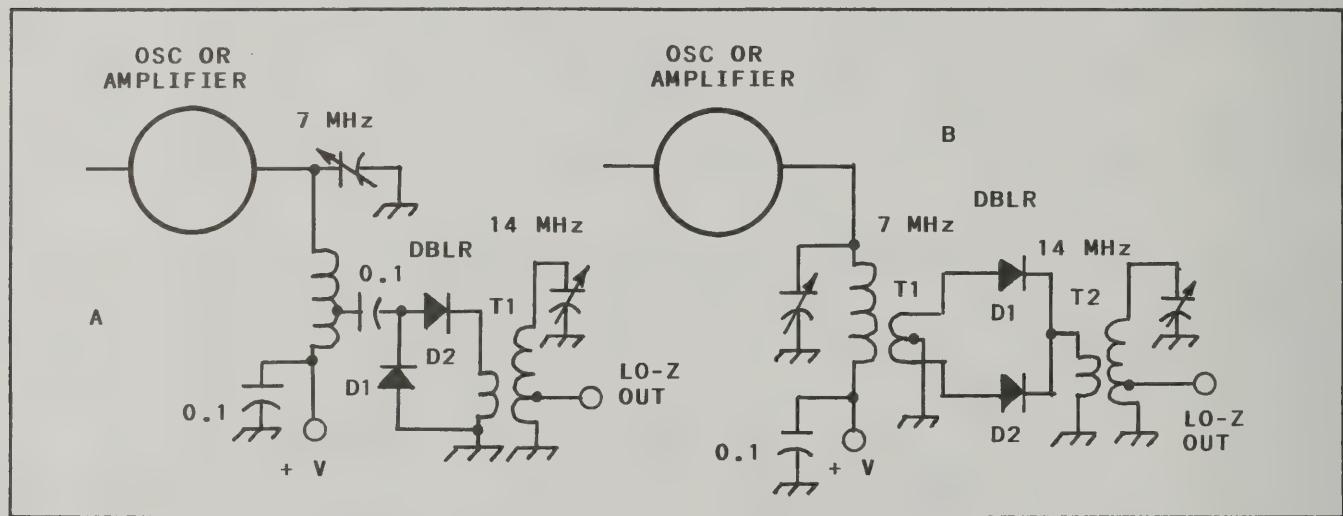


Fig 3-7 -- Two examples of low-level diode frequency doublers. Type 1N914 high-frequency silicon diodes are suitable for these circuits, although 1N34A germanium diodes may also be used. Circuit A shows a frequency doubler that is also a voltage doubler. If sufficient drive is available for D1 and D2, the harmonic output will be high. T1 is tuned to the second harmonic of the diode input frequency. Circuit B shows a balanced diode doubler circuit. T2 is tuned to the second harmonic of the diode input frequency. If the loaded Q of T2 is high it is possible to obtain up to 40 dB of rejection for the 7-MHz energy. The 21-MHz energy will be suppressed in a like manner.

The circuits depicted in Fig 3-7 are useful when we build transmitters and local-oscillator circuits where frequency multiplication is required. Output from the doublers can be boosted by adding a class A broadband amplifier at the doubler output. Note that the input to the diodes in both examples is low impedance. Impedance values from 50 to 200 ohms are generally satisfactory when tapping the diodes on a tuned circuit or when using link coupling to the diodes, as in Fig 3-7B. The center tap for the T1 secondary needs to be at the electrical center of the winding. Alternatively, you may use a bifilar link on T1 to improve the balance.

Zener Diode Usage

Zener diodes are used as voltage regulators and to establish a fixed-voltage reference in other types of regulator circuits. They are used also as protection devices that limit the peak voltage in a particular type of circuit. Still another use for Zener diodes is for series gating where a voltage is not desired until it reaches the rated conduction voltage of the Zener diode used. In this application the diode functions as a go-no-go device.

Zener diodes are available with power ratings from 400 mW to many watts. They are manufactured for conduction voltages from 3.9 to 180. A listing of the available voltages and wattage ratings is given in **The ARRL Electronics Data Book**. Fig 3-8 contains examples of how we may use Zener diodes for their most common applications.

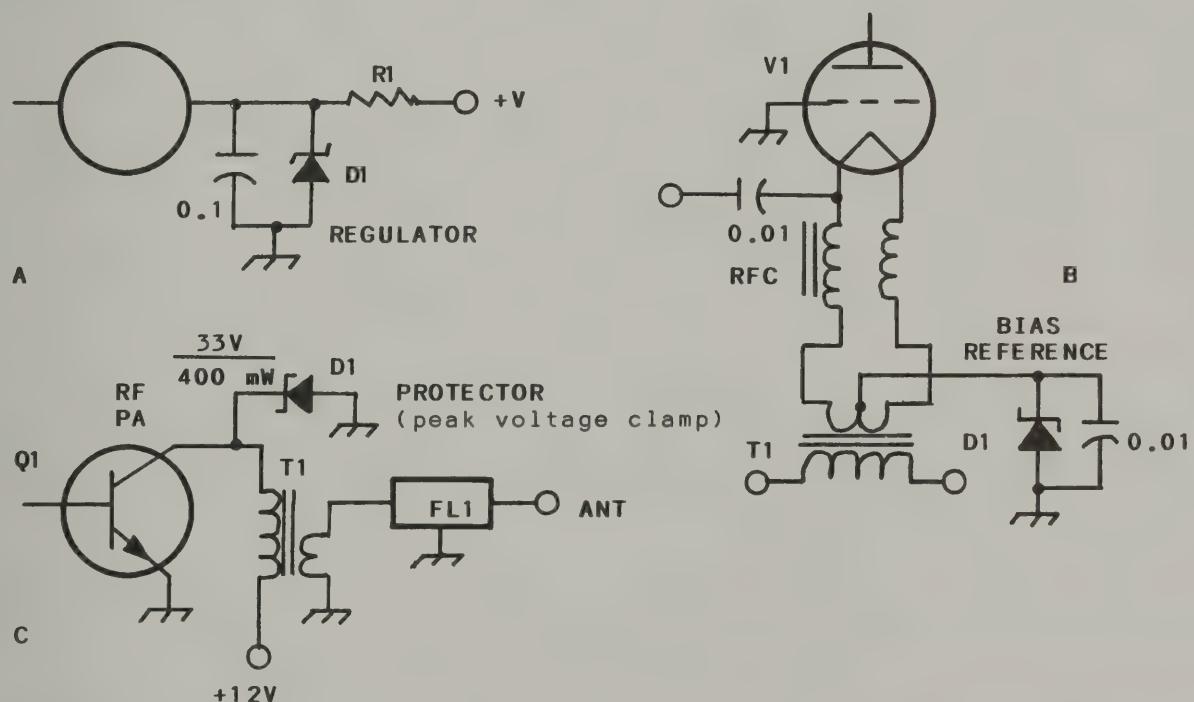


Fig 3-8 -- Three examples of how Zener diodes are used. The circuit at A has D1 in use as a shunt voltage regulator. See text for determining the proper value for R1. Circuit B shows a grounded-grid RF power amplifier that uses a high-wattage Zener diode (D1) to develop the bias for V1. The grid thus becomes negative with respect to the cathode. At C is an example of how to use a Zener diode to protect Q1 from high SWR or voltage peaks that can occur during periods of self-oscillation. D1 also protects against voltage spikes that may occur on the dc supply line.

The resistance and wattage rating of R1 at Fig 3-8A can be calculated by means of Eq. 3-1 and Eq. 3-2.

$$R1 = \frac{E_{in(min)} - EVR}{I_L + 0.1I_L} \quad \text{Eq. 3-1}$$

where R is in ohms, E_{in} is the supply voltage, E_{VR} is the rated Zener-diode voltage and I_L is the current taken by the circuit to be regulated. Typical current flow through the Zener diode for this circuit is approximately 18 mA for good regulation. I_L in the equations is expressed in amperes.

$$P_D(\max) = \left[\frac{E_{in(\max)} - E_{VR}}{R} - I_L \right] E_{VR} \quad \text{Eq. 3-2}$$

where P_D is in watts and the remaining terms are the same as in Eq. 3-1.

Use care when selecting a Zener diode for D1 in Fig 3-8C. The larger the diode geometry (inner workings) the greater the internal capacitance. Too great a capacitance value at the Q1 collector can present an impedance that can disturb the T1 impedance match (caused by the diode X_C). Furthermore, the unwanted C can bypass the RF energy to ground and render the stage inefficient. This effect is the most pronounced in high-Z circuits and at the upper end of the HF spectrum. I have found that most 400-mW Zener diodes are fine up to 30 MHz. Many 1-watt diodes are suitable also, especially below 14 MHz.

The Zener diode chosen for D1 of Fig 3-8B must be capable of passing the cathode current of V1 without overheating. It should be mounted on a heat sink in order to enhance cooling. A 50-W Zener diode is suitable for circuits that provide the maximum legal Amateur Radio output power.

Zener diodes may be used in series to obtain a specific regulated-voltage value. The diodes need not have the same voltage rating if this is done. In fact, it is sometimes convenient to use two or more Zener diodes of different values (series connected) to provide various regulated voltage pick-off points. You may also place a silicon diode in series with a Zener diode to increase its voltage barrier by 0.7 V. For example, if you have a 9.1-V Zener diode, but wish to create a 10-V regulated source, you may use a silicon rectifier diode in series with the Zener diode to obtain 9.8 volts, regulated.

An LED is useful as a reference diode. It conducts at 1.5 volts, and may be used to establish bias by placing it in series with a transistor emitter lead that goes to ground. It can also be used to provide a 1.5 volt, low-current regulated voltage.

Another hint and kink I wish to pass along to you is that you can use a small-signal NPN transistor as a Zener diode. This may be done by grounding the collector and applying +dc voltage to the emitter. The base is connected to the collector when this is done. Most transistors of the 2N2222 and 2N3904 type will yield a regulated voltage of approximately +8 when this is done.

PIN Diodes

PIN diodes behave as voltage-variable resistors. They are used mainly at VHF and above to function as band switches or as short circuits across antenna ports or waveguides. The Motorola MPN3700, for example, has a turn-on dc resistance of only 0.7 to 1.3 ohms. The resistance value is proportional to the diode forward current in mA. The greater the current the lower the resistance. This diode can handle up to 200 mW of power dissipation. The diode capacitance is 1 pF and it

has a reverse recovery time (t_{rr}) of 300 ns. Maximum safe reverse voltage (VR) is 200. PIN diodes are not widely used by amateurs, except for those hams who work at microwave frequencies. PIN diodes may be used also as AGC devices in VHF receivers if they are used at the antenna-input circuit of the receiver.

Schottky and Hot-Carrier Diodes

Schottky or hot-carrier diodes are excellent for high-speed solid-state switching. They have a lower barrier voltage than is available from silicon switching diodes. They are point-contact diodes rather than being junction diodes. The point-contact structure is similar to that of the old crystal detector that used a piece of galena crystal and a cat's-whisker contact electrode. This structure minimizes internal capacitance, which in turn shortens the diode switching time. These diodes are excellent devices for use in mixers. A popular hot-carrier diode is the Hewlett-Packard HP-2800.

Rectifier Diodes

Diodes that are used in power supplies to convert ac to dc energy are covered in detail in **The ARRL Handbook**. Most amateurs are familiar with their selection and use. It would consume page space unnecessarily if I were to include a treatise on this subject here. Suffice to say that like other diodes, the user must pay close attention to the diode current rating and PRV (peak reverse voltage) specifications in accordance with the operating voltage of the power supply and the load current presented by the equipment for which the power supply is designed. These subjects are covered in **The ARRL Handbook**.

IC CIRCUIT APPLICATIONS

Integrated circuits offer us the opportunity to pack a lot of circuitry into a rather small area. All is not milk and honey when we use these modern components, because circuit-board layout becomes more difficult than with discrete parts, and we have little control over what takes place within many of these chips. A circuit that has discrete components is accessible when we wish to alter the operating voltages and currents, add capacitors or make other minor circuit changes. An IC may have dozens of internal transistors, resistors, diodes and capacitors. We can only change the operating parameters at the pins of the IC. Circuit samples that appear on the data sheets for a number of ICs are often too basic to permit top performance, and it is not uncommon to find that the operating voltages and bias values can be improved upon by the user. An amateur with an inquiring mind and the courage to experiment can often make an IC exceed its published specifications by making cautious changes in signal levels and IC current. I encourage you to explore this possibility when you desire optimum performance.

Audio ICs & Applications

Integrated circuits may be used as audio preamplifiers and as audio power-output stages in receivers. The popular 741 op amp (operational amplifier) is excellent for use as a headphone driver, low-level audio amplifier and as an RC active audio filter. These low-cost ICs can be set for voltage gains (AV) from unity to +40 dB by selection of the feedback-resistor value. Furthermore, you can purchase an op-amp IC that contains one, two or four op amps in a single IC package. This makes it possible to use one section as an audio preamp while using the remaining two or three op amps as sections of an RC active audio filter. If we were to build

a similar circuit with discrete components it would require dozens of transistors and resistors that are already present in the IC!

Op amps tend to be noisy compared to low-noise audio transistors. They generate what is known as "pop corn" noise. This is seldom a problem if the op amp audio stage is preceded by a low-noise bipolar transistor or FET. Low-noise op amps are available for circuits in which a low-noise preamp is not used. They are called "biFET" op amps. The input devices in these chips are FETs, which do not generate the level of noise that is created by op amps that have bipolar transistors in the input circuit of the IC. An example of a biFET op amp is the TL080. The cost is slightly greater than for a 741 op amp, but they are worth considering when you design a circuit.

Fig 3-9 contains practical examples of ICs as they may be applied in low-level audio service.

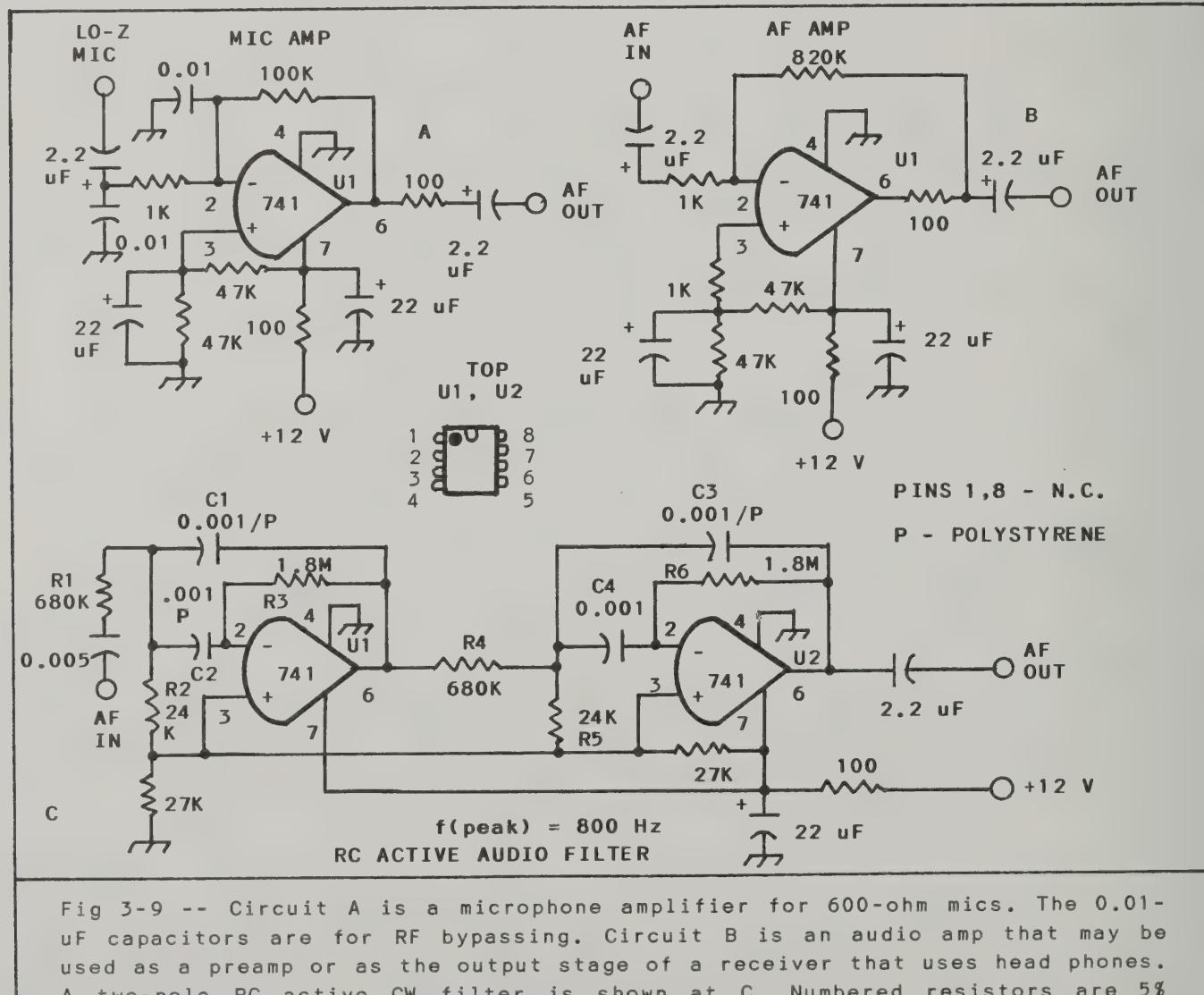


Fig 3-9 -- Circuit A is a microphone amplifier for 600-ohm mics. The 0.01-uF capacitors are for RF bypassing. Circuit B is an audio amp that may be used as a preamp or as the output stage of a receiver that uses head phones. A two-pole RC active CW filter is shown at C. Numbered resistors are 5% types. C1 through C4 should be closely matched in value.

The op amps in Fig 3-9 may be replaced with TLO-81 biFET chips in the interest of reduced noise. The pin-out is the same as for a 741. You may wish to use a dual op amp for U1 and U2 in Fig 3-9C. A 747 op amp may be used for this purpose. This two-pole RC active CW filter requires close matching of the numbered resistors and capacitors in order to have the peak response occur at one frequency. C1-C4 must be a high-Q type, such as polystyrene or mylar. The resistors in all three Fig 3-9 circuits are 1/4-W carbon composition or carbon film types. The polarized capacitors are electrolytic or tantalum. Narrower skirts for the filter response curve can be obtained by adding a third pole to the filter. Place two U1 stages in series and follow them with the U2 circuit to develop a three-pole filter. The values of the numbered components may be changed to obtain peak responses at frequencies other than 800 Hz. Equations for RC active filters are presented in *The ARRL Handbook* and in *Solid State Design for the Radio Amateur*.

An RC active filter should be used at low audio levels in order to prevent it from being overloaded. I prefer to insert them directly after the 1st audio amplifier in my receivers. These filters may be designed to have a low-pass, high-pass or bandpass response. The Fig 3-9C circuit has a bandpass response, which enables it to reject low and high frequencies, respective to the center frequency. The filter has a Q of 5 and a gain of 2.

IC Audio Power Amplifiers

Most audio power ICs are designed for use with 4- or 8-ohm speakers. They do not require an input or output transformer. This minimizes cost and PC-board area. Fig 3-10 shows practical circuits for audio power amplifiers.

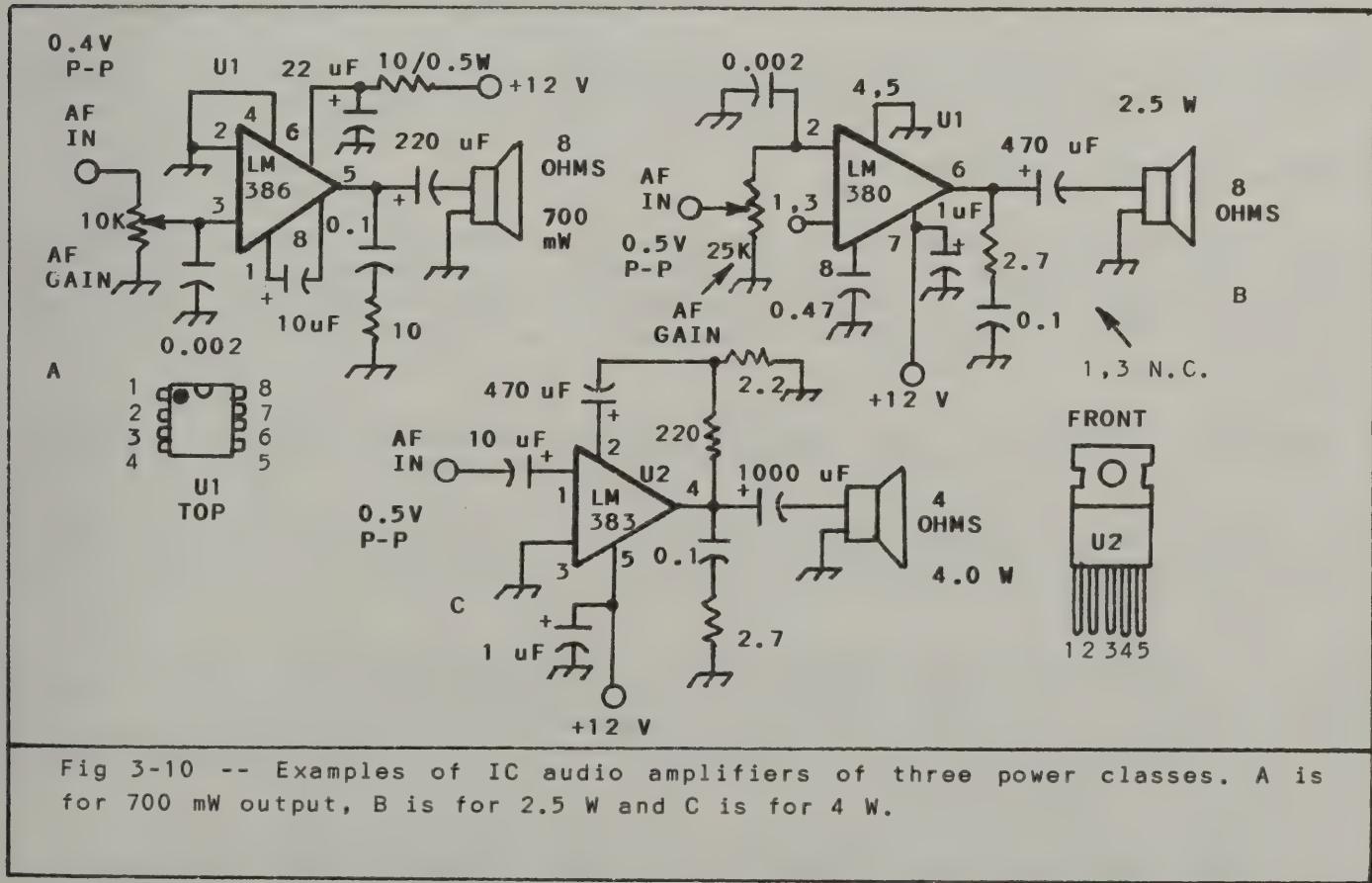


Fig 3-10A shows an excellent audio-amplifier choice in terms of low cost and modest power consumption. This amplifier is able to provide normal room volume with 3- or 4-inch speakers. It is a popular circuit for hams who build QRP and portable equipment. The 0.1- μ F capacitor and 10-ohm resistor to ground from pin 5 of U1 discourage self-oscillation. Most IC audio amplifiers have a tremendous amount of gain in a very small package, and this encourages instability. The LM386, as shown, has a gain of 200!

Fig 3-10B shows a more powerful IC audio amplifier. It can deliver 2.5 watts of output with a supply voltage of 12. It can provide 10 watts of output power into a 4-ohm load if the operating voltage is increased to 22, but these are the safe maximum ratings for the IC. Maximum peak current for the LM380 is 1.3 amperes. The circuit shown is for the LM380N-8. This chip is available also in a 14 DIP package (LM380N).

The amplifier in Fig 3-10C will deliver 7 watts of output power into a 4-ohm load when the supply voltage is 16. Idling current at 12 volts is 40 mA. Peak repetitive output current for the LM383A is 3.5 A. This is a good audio IC to use in a high-performance receiver, because normal room volume is less than 1 watt. This allows the IC to operate well below the level where distortion becomes apparent to the ear.

All bypass capacitors for the circuits in Fig 3-10 must be located as close to the related IC pins as practicable. This is important if the bypassing is to be effective toward preventing self-oscillations. A heat sink is required for the Fig 3-10C amplifier. I suggest that you affix a small heat sink to the LM380N-8 also. This may be done by attaching the sink with epoxy cement.

You may use 4- or 8-ohm speakers with each of the amplifiers in Fig 3-10. Circuits B and C will produce greater output power when a 4-ohm speaker is used.

Universal IC Building Blocks

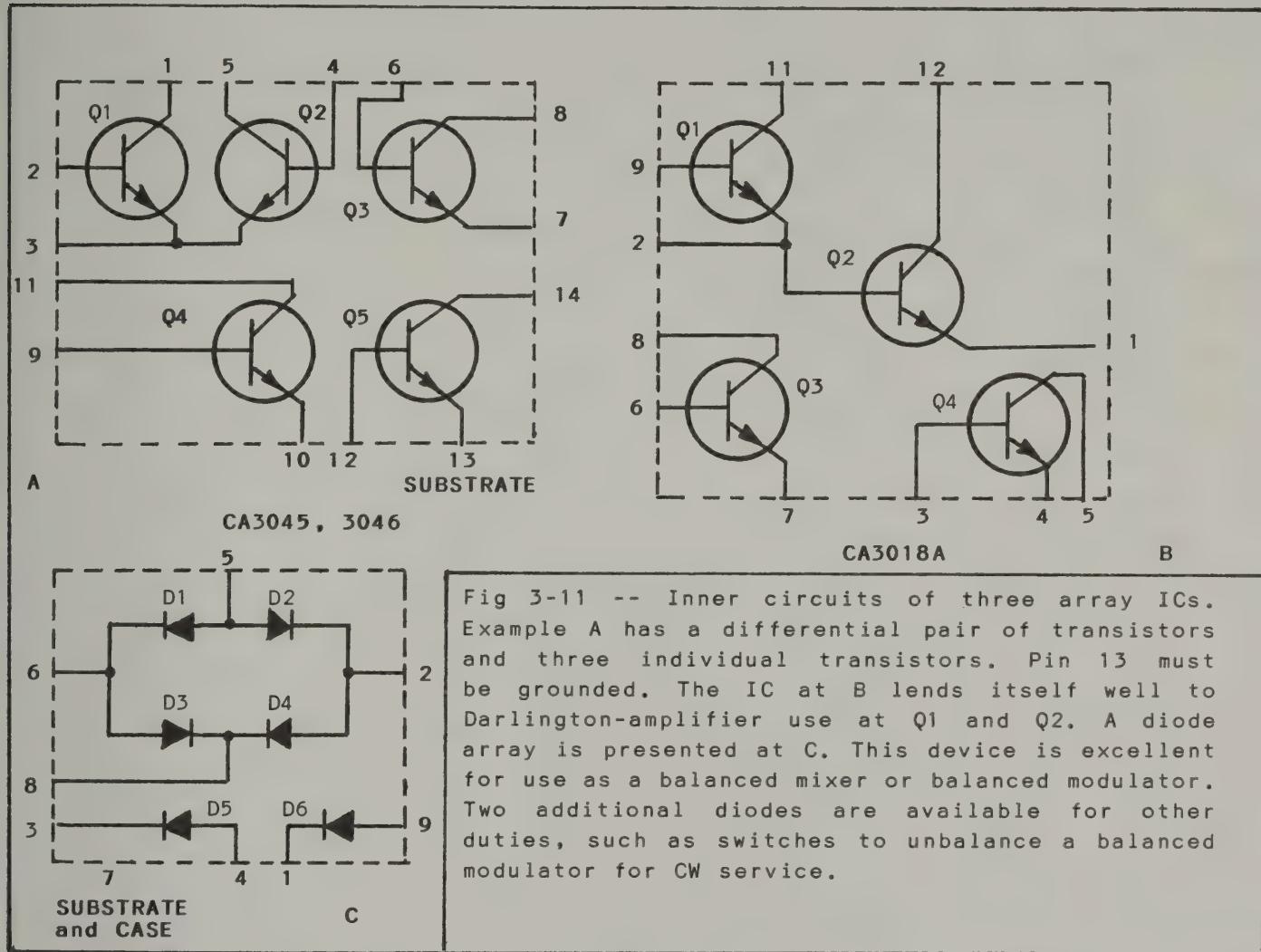
Transistor-array ICs appear tailor made for experimenters. These devices lend themselves to all manner of amateur circuits from receivers to transmitters. They contain individual diodes or bipolar transistors that are accessible at the various IC pins. Most transistor arrays are suitable for use from low frequency well into the VHF spectrum.

A particular advantage we enjoy when working with arrays is that each transistor on the IC substrate is formed at the same time, and under the same manufacturing conditions. This virtually ensures that the transistor operating characteristics are identical or nearly so. This feature is of particular value when we build balanced circuits, such as mixers, differential amplifiers, balanced modulators and push-pull RF amplifiers.

An entire direct-conversion receiver can, for example, be built around an RCA CA3045 or CA3046 array. This chip contains the circuit shown in Fig 3-11A. Another interesting transistor array is the RCA CA3018A (Fig 3-11B). It can be made into a tiny QRP transmitter, to give but one example. The utility of these ICs is limited only by your imagination.

You can make your own doubly balanced active mixer or balanced modulator by using the RCA CA3026, CA3049 or CA3054 IC arrays (Fig 3-12). The CA3019 is a diode-array IC that is ideal for use as a doubly balanced passive mixer or balanced

modulator. The inner workings of the aforementioned ICs are shown in Fig 3-11.



The ICs above offer some interesting circuit possibilities. For example, Q1 and Q2 in the CA3045 IC can be used as a singly balanced detector for a direct-conversion (DC) receiver. Q4 might be used as the tunable oscillator or variable frequency beat oscillator (VFO). Q3 and Q5 could be used as audio amplifiers for a pair of headphones. The CA3018A at B of Fig 3-11 could employ Q3 as a VFO, Q4 as a buffer and the Q1-Q2 direct-coupled pair as a VFO buffer-amplifier.

The diode array at C of Fig 3-11 provides the foundation for a doubly balanced mixer or balanced modulator. External broadband ferrite transformers can be added for coupling to and from the circuit.

Other types of transistor-array ICs are available from several semiconductor manufacturers. These chips make possible the development of the old-time comic-strip Dick Tracy wrist radio! Certainly, they should have a place in your experimenter's bag of tricks.

There are three RCA integrated circuits that contain the twin differential amplifiers shown in Fig 3-12. They are the CA3026, CA3049 and CA3054. The pin-out for the CA3026 is different from that of the other two ICs. These chips are excellent

for active mixers or balanced modulators when we need a doubly balanced circuit. The principle advantage when using active circuits is that we obtain a conversion gain rather than a loss. Gains of up to 15 dB are common, whereas a doubly balanced diode-ring mixer has a typical conversion loss of 7-8 dB.

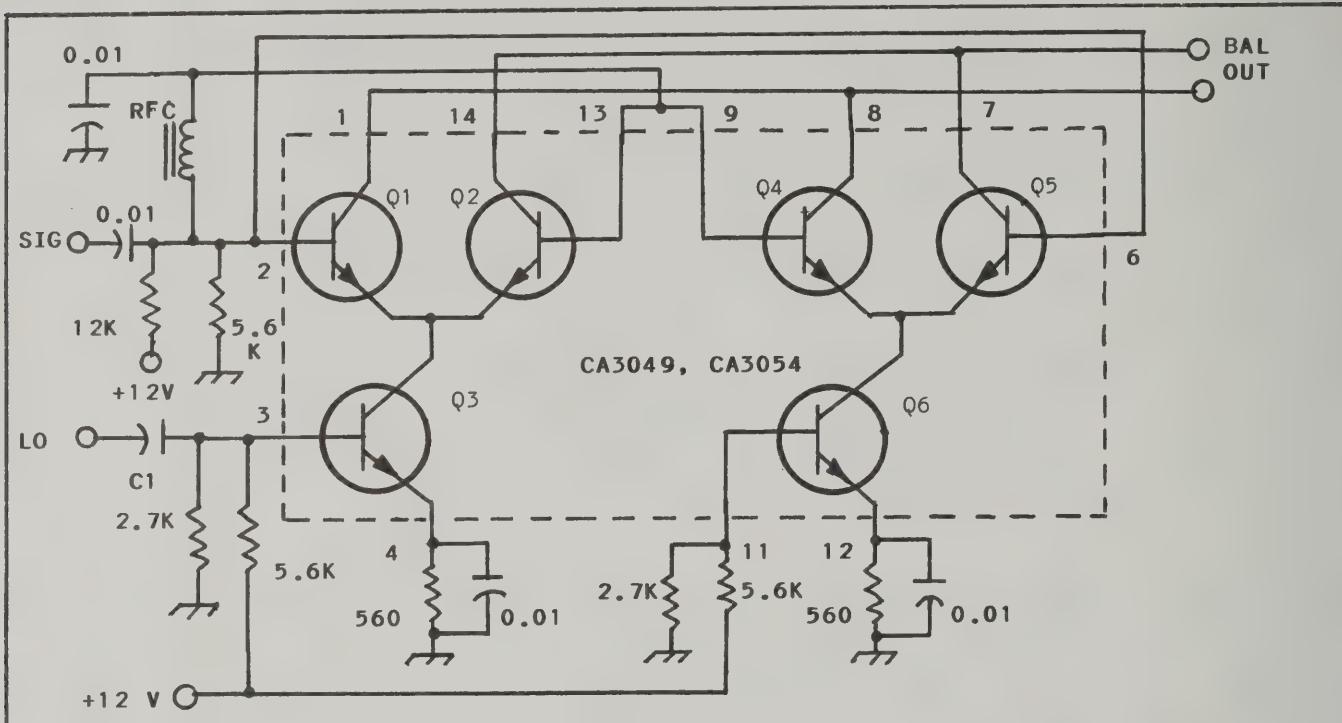


Fig 3-12 -- Circuit for a doubly balanced mixer or balanced modulator using a CA3049 or CA3054 IC. The dashed lines divide the inner working of the IC from the external components. Two differential transistor pairs (Q1, Q2 and Q4, Q5) are cross-coupled as shown to form a doubly balanced circuit. Each differential pair has its own current source transistor (Q3 and Q6). LO injection is applied to the base of the Q3 current-source transistor. The RF choke isolates the signal input from the bypassed Q2 and Q4 bases. A 1-mH choke is suitable for HF band operation. The value of C1 is chosen to ensure 4-5 volts P-P of LO injection.

When we use an IC of the type illustrated in Fig 3-12 it is essential that we strive to obtain a symmetrical layout for the PC board. The circuit-board conductors should be as short and direct as practicable. Wide PC-board foils help to minimize stray inductance. These precautions aid the overall circuit balance. The Fig 3-12 circuit is excellent also as a product detector and as a front-end circuit for a DC receiver.

IC Audio Oscillators

Experimenters have frequent need for audio oscillators that can be used as tone generators for monitoring one's CW sending. ICs lend themselves well to oscillator service. These circuits may be used also as function generators for testing Hi-Fi and other audio circuits. Various kinds of IC, such as timers, op amps and audio amplifiers may be used as oscillators. When we use them as sidetone oscillators we can key them on and off by simply breaking the Vcc supply line to them. The output level is generally set by means of a potentiometer that is adjusted for comfortable headphone or speaker volume. Fig 3-13 shows two IC audio oscillators.

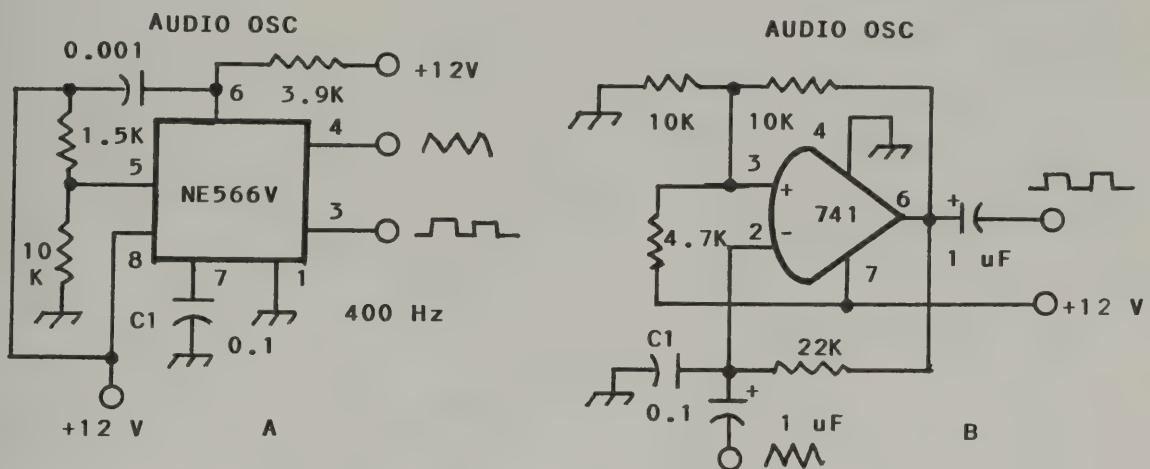


Fig 3-13 -- Examples of IC audio oscillators that may be used as function generators (triangular and square waves) or keyed sidetone oscillators. C1 at A and B is selected for the audio tone desired.

Both of the ICs in Fig 3-13 are 8-pin DIP types. If you do not like to listen to the harsh square or triangular waves you may add an audio filter after the oscillator to "launder" the oscillator waveform. A filter will provide a sine wave output. The circuit of Fig 3-13B can be built around a dual op amp, such as the 747. One section would then serve as the tone generator while the remaining op amp would be used as a single-pole RC active bandpass filter. If you decide to use this method for obtaining a softer output waveform, make certain that the center frequency of the filter matches the output frequency of the oscillator.

ICs as RF and IF Amplifiers

Integrated circuit RF amplifiers offer high gain in a small package. Another advantage associated with their use is that most ICs that are designed for IF-amplifier service have provisions for the application of AGC voltage. A vast number of these ICs are on the market, but we will focus on only a few of the more popular amplifier chips that are used by amateurs.

Although these ICs can be used as RF amplifiers in the front ends of communications receivers, I don't recommend that they be used for this purpose above 7 MHz. This is because most IC RF amplifiers are more noisy than FETs and bipolar transistors that are designed for low-noise operation. Fig 3-14 contains practical examples of ICs in IF-amplifier service.

Note that the AGC control voltage is different for each circuit. Maximum amplifier gain occurs at +9 volts for the CA3028A, whereas maximum gain for the MC1350P is obtained at the low end of the control-voltage range.

Typical gain for an integrated circuit IF amplifier is 25-40 dB. The magnitude of the gain is dependent upon the ratings of the chip, input-output impedance matching and the biasing of the IC. Since the potential gain is high, we must use care in the circuit layout (short, direct leads), and we need to install our bypass capacitors as close to the IC pins as practicable. Strive toward the best possible isolation of the IC input and output circuits in the interest of stable operation.

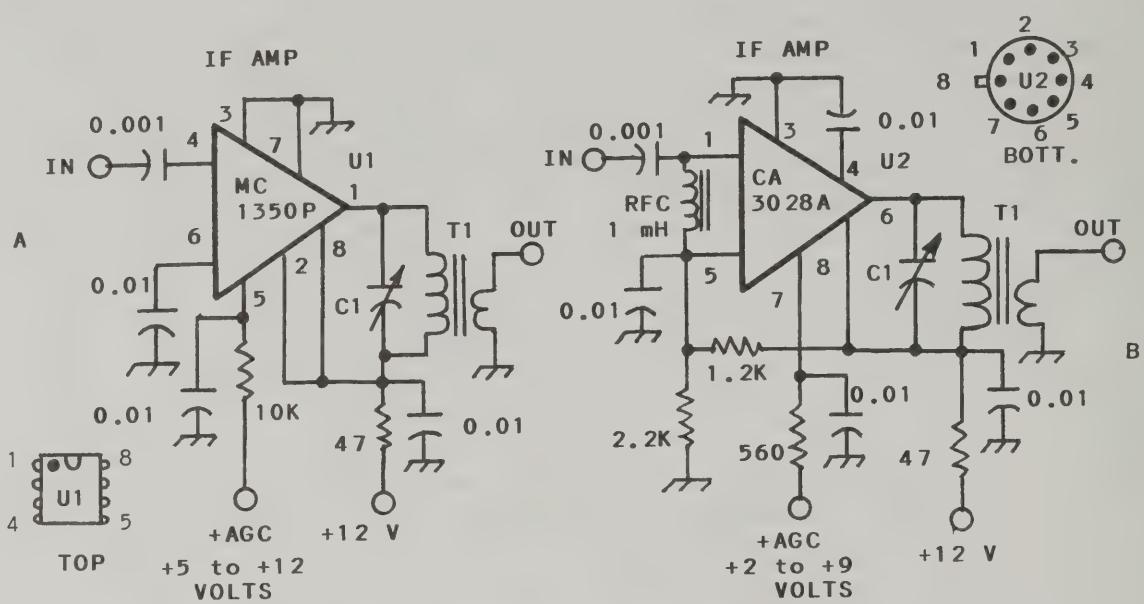


Fig 3-14 -- Practical examples of two IF amplifiers that employ ICs. Circuit A contains a low-cost Motorola MC1350P. AGC control voltage is +5 to +12, with maximum amplifier gain at +5 V. C1 and T1 in both examples are tuned to the IF. Approximate input Z for both circuits is 1K ohms. Output impedance is on the order of 4K ohms. Circuit B uses an RCA CA3028A IC. AGC voltage for this device is +2 to +9, with maximum gain occurring at +9 V. Both amplifiers can provide up to 40 dB of gain.

The IF amplifiers in Fig 3-14 are generally used in pairs. That is, two stages are used in cascade in order to obtain 80 dB of overall gain. The AGC action is more effective if two stages are used, and with AGC voltage applied to each of them. Circuit A can be used with the MC1590G, which is considered a high-performance IF amplifier IC. The pin-out is different for these two devices. Examples of how cascaded MC1590G and CA3028A IF amplifiers are given in practical receiver circuits that may be found in *Solid State Design for the Radio Amateur* (ARRL), pages 136 and 228.

It is practical to use but one of the stages in Fig 3-14 when you build a simple receiver. The AGC control voltage may be replaced with a manually controlled voltage of the same range. This modification permits the use of manual IF-gain control. An example of this method is on page 107 of the above-referenced book.

IC Product Detectors

Earlier, we treated the similarity between mixers and product detectors. In effect, we apply an input signal and an LO signal (BFO in this instance) and extract audio-frequency energy rather than an RF (IF) signal. In a similar fashion we may compare a product detector to a reverse balanced modulator, since both devices are associated with an RF signal, a local oscillator and audio energy. The audio output from a product detector (PD) is the product of the two input signals, and hence the name. For example, if we apply an IF of 455 kHz to the PD and the BFO is on 456 kHz, we will obtain a 1-kHz (1000 Hz) audio tone at the output port of the PD. The same 1-kHz tone will occur if the BFO is on 454 kHz.

As is the situation with active mixers, an active (requiring an operating voltage) PD has gain rather than conversion loss. This is an advantage when we wish to reduce the number of IF or AF stages in a bare-bones type of receiver. The overall gain of a practical receiver should range from 80 to 100 dB (front end to audio-output) to ensure adequate weak-signal sensitivity for headphones or a speaker. This is not true if we are dealing mainly with signals that are S3 or greater. Fig 3-15 shows practical circuits for product detectors.

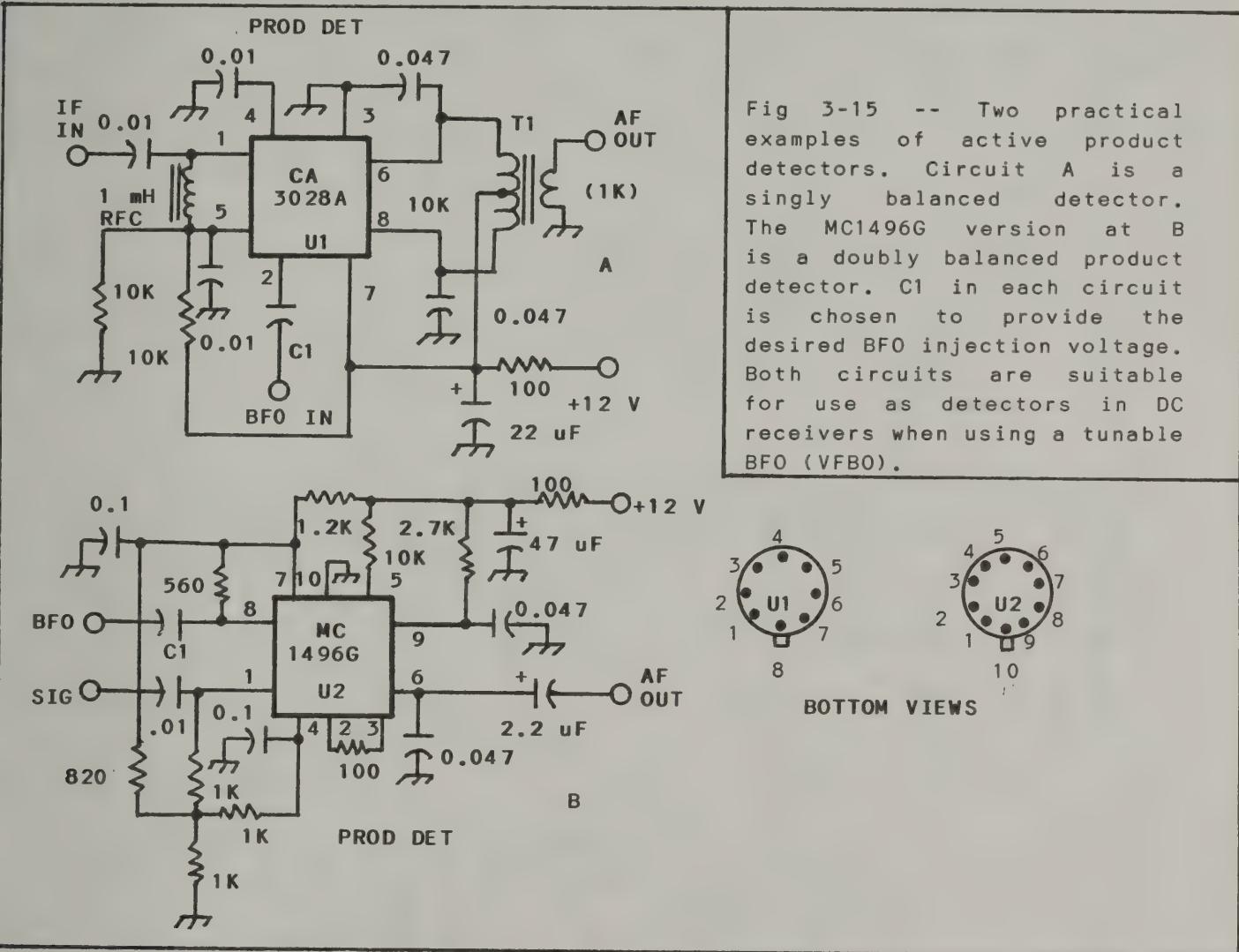


Fig 3-15 -- Two practical examples of active product detectors. Circuit A is a singly balanced detector. The MC1496G version at B is a doubly balanced product detector. C1 in each circuit is chosen to provide the desired BFO injection voltage. Both circuits are suitable for use as detectors in DC receivers when using a tunable BFO (VFBO).

Fig 3-15A show how to use the familiar RCA CA3028A as a singly balanced PD. T1 is a miniature audio transformer with a 10K-ohm center-tapped primary and a 1K-ohm secondary. The bypass capacitors at pins 3 and 8 can be chosen to provide a band-pass audio response with the primary inductance of T1. These capacitors also help to roll off high-frequency response and reduce audio hiss. BFO input voltage should be on the order of 1 to 1.5 V RMS.

The product detector in Fig 3-15B is doubly balanced. It uses RC output coupling, but T1 of circuit A may be used at the U2 output by connecting it to pins 6 and 9. Both product detectors in Fig 3-15 are excellent performers when used as front ends of DC receivers. A high-Q tuned input circuit is connected to the IF input ports for DC-receiver use. BFO injection to U2 of Fig 3-15 is 300 mV RMS.

IC Mixers

Good receiver performance is dependent in large measure upon the dynamic range of the receiver front-end circuits. The diode-ring mixers in Fig 3-1 offer excellent dynamic range (DR). However, a number of balanced-mixer ICs are capable of providing high dynamic range if the input-signal levels are not excessive, as specified by the manufacturer. If an RF amplifier precedes an IC mixer, it must also have the ability to accommodate large signals without degraded IMD (intermodulation distortion) performance. Proper gain distribution is vital to good performance, especially at the receiver front end. Too little gain can result in a poor receiver noise figure, and too much gain leads to poor dynamic range. The correct LO injection level to a mixer plays a vital role also, respective to DR and mixer gain. The subject of receiver dynamic range is covered in **Solid State Design for the Radio Amateur** and in **The ARRL Handbook**. Therefore, I will not treat the subject here.

Fig 3-16 illustrates IC mixers with assigned, practical component values. The mixers shown are among the most popular ones used by amateurs, although a number of other ICs may be used for this application.

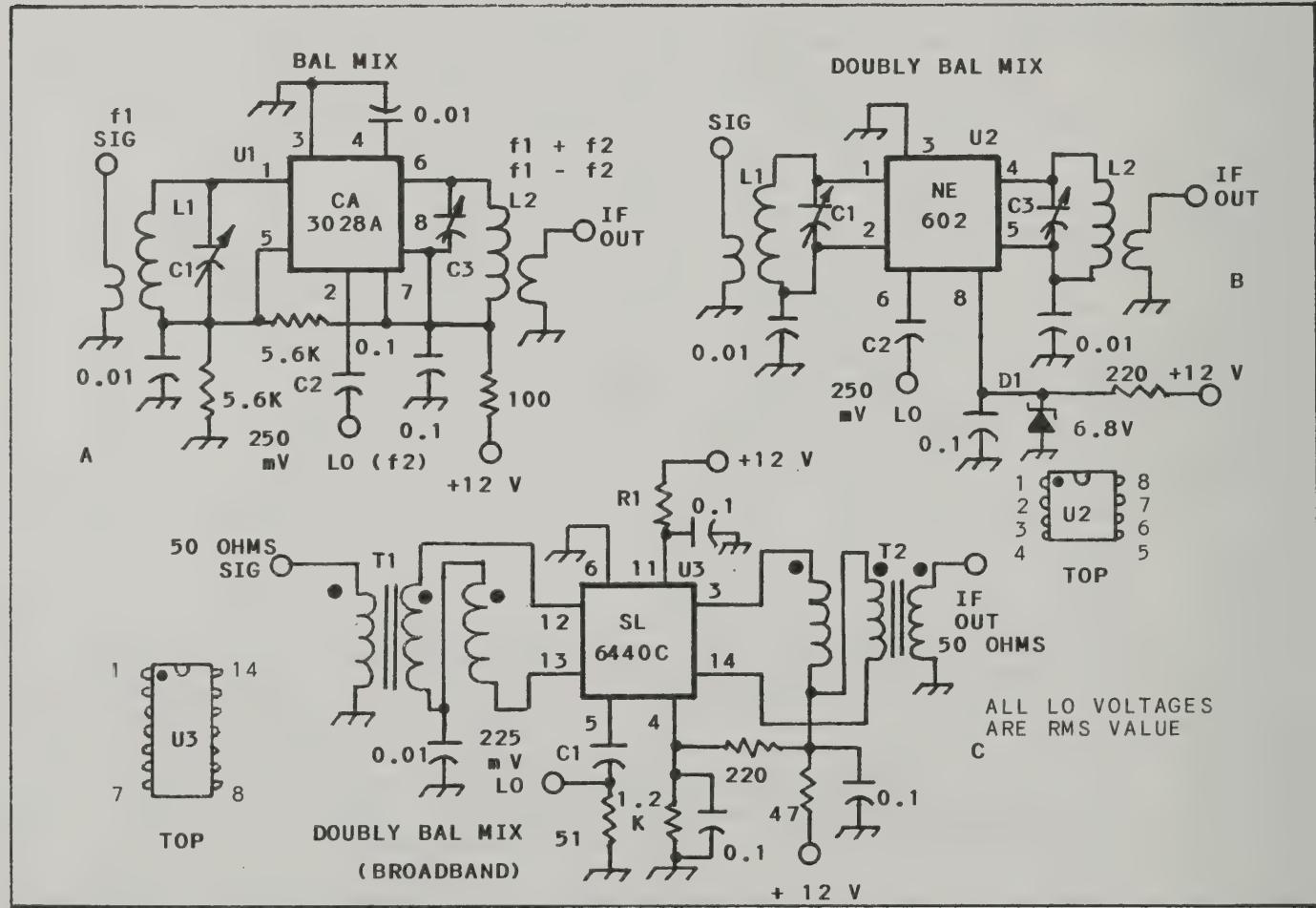


Fig 3-16 -- Three popular IC mixers. Circuit A is singly balanced. Circuits B and C are doubly balanced. The broadband mixer method shown at C may be applied to circuits A and B. See Fig 3-15 for the CA3028A pin out. The series capacitor at each IC LO pin is chosen to establish the mV injection value listed, which should appear at the related IC pin.

Fig 3-16A shows a singly balanced mixer that uses an RCA CA3028A IC. The input and output tuned circuits are arranged in a single-ended format, but both resonators may be connected in a push-pull fashion to improve the circuit balance. The broadband arrangement at C of Fig 3-16 may be used for circuits A and B to accomplish this, or tuned circuits with L1, L2 center taps may be employed. If tuned circuits are used as shown, C1 and L1 are resonant at the signal frequency, while C3 and L2 are tuned to the intermediate frequency.

The MC1496G of Fig 3-15B may be used also as a doubly balanced mixer with the parts values indicated. Pin 6 becomes the IF output of the IC. The 0.047- μ F capacitor is eliminated, as is the 2.2- μ F coupling capacitor. Pin 6 is routed to a tuned circuit (as in Fig 3-16A), with the bypassed end of the tuned circuit connected to pin 9 of the MC1496G.

A doubly balanced mixer with very few external components is seen at B of Fig 3-16. The Signetics NE602 has become a popular IC among experimenters because it not only functions well as a mixer, but has provisions to function also as a local oscillator. Pins 6 and 7 are used for this purpose. The external components required are used to create a crystal-controlled or LC type of oscillator. A typical circuit is presented in Fig 3-17. The LO section will operate to 200 MHz. Mixer conversion gain may reach 18 dB at 45 MHz. The noise figure (NF) is specified as 5 dB at 45 Hz.

Fig 3-16C shows a high dynamic-range Plessey SL6440C mixer. The broadband input and output transformers contain 15 trifilar turns of no. 30 enamel wire on an Amidon T-37-43 ferrite toroid (850 μ u). The polarity of the windings must be as indicated by the black dots. R1 is selected to cause the IC to draw 13 mA of current. The input intercept for this chip is rated at +31 dBm. Conversion gain is 0 dB and output intercept (OI) is +31 dBm. A post-mixer amplifier may be added to compensate for the unity conversion gain.

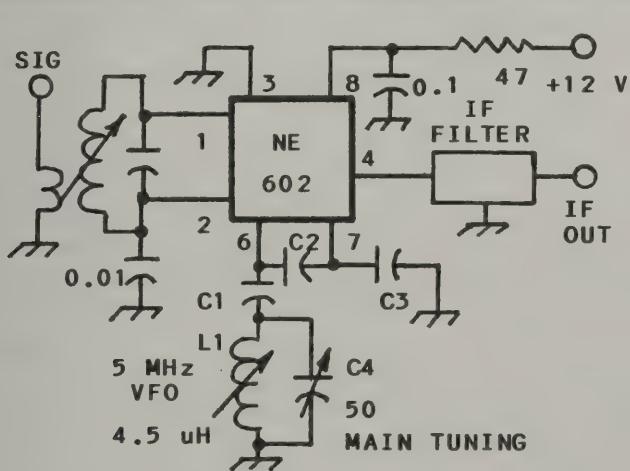


Fig 3-17 -- The Siliconix NE602 mixer may be used as a mixer-osc as shown here. A 5-MHz LO example is given. C1, C2 and C3 are NPO capacitors. C1 is 470 pF and C2, C3 are 560 pF. An IF filter may be connected to pin 4 in lieu of an IF tuned circuit. A 9-MHz filter is appropriate if the signal freqency is 4.0 MHz and the LO is at 5 MHz. The same LO and intermediate frequency also permits a 14-MHz signal frequency.

In Fig 3-17 pin 6 is the base of the internal oscillator transistor and pin 7 is the emitter. The circuit above is for a Colpitts oscillator. This use of the NE602 is ideal when we wish to build simple DC and superheterodyne receivers. It is better to use an outboard LO with buffering for high-performance receivers. Oscillator harmonics can cause performance problems when the LO transistor is on the same substrate as the remainder of the IC circuitry.

ICs as Audio Filters

The IC technology has made it possible for us to construct audio filters that are small and lossless. Gone are the days when we had to utilize bulky inductors and capacitors to build a low-pass, high-pass, band-pass or notch filter. We can cause an active RC (resistance-capacitance) filter to have unity or greater gain. We can control the filter Q by means of component selection, which is always a problem when we design LC types of filters. OP amps (operational amplifiers) are used for RC active filters. The generic 741 op amp is often the choice of amateurs. The op amp noise is lower when a biFET such as the TL080 is used in an RC active filter. This is an important consideration if the filter is used in the low-level section of an audio amplifier, since the filter may establish the overall noise figure of the amplifier chain. This may be equated to the use of a low-noise RF amplifier ahead of a mixer.

Fig 3-18 contains circuits for a low-pass and a notch filter for use at audio frequency. The notch filter is adjustable and has a null depth of some 40 dB. The low-pass filter has but one pole, as shown. Normally, two or three identical stages are used in cascade to enhance the high-frequency rolloff beyond the -3 dB point on the filter response curve.

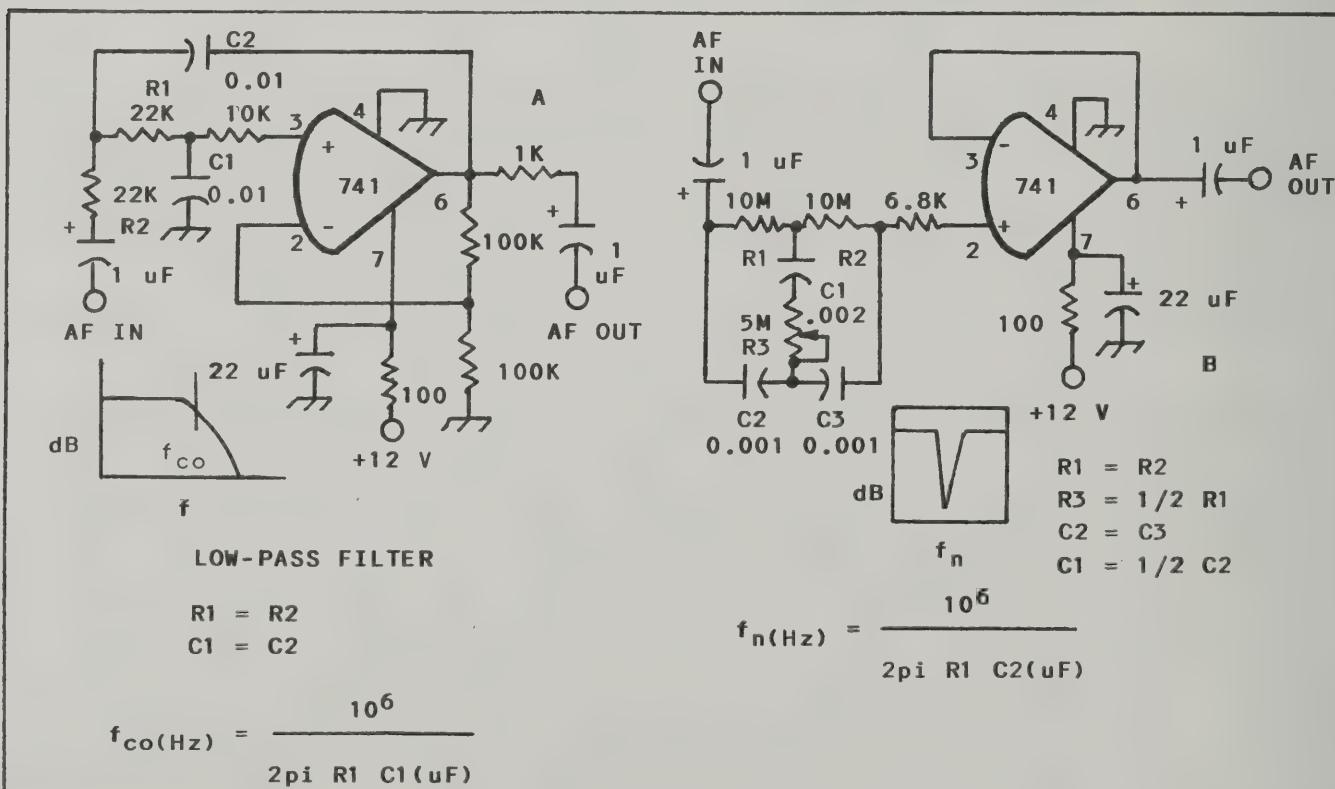
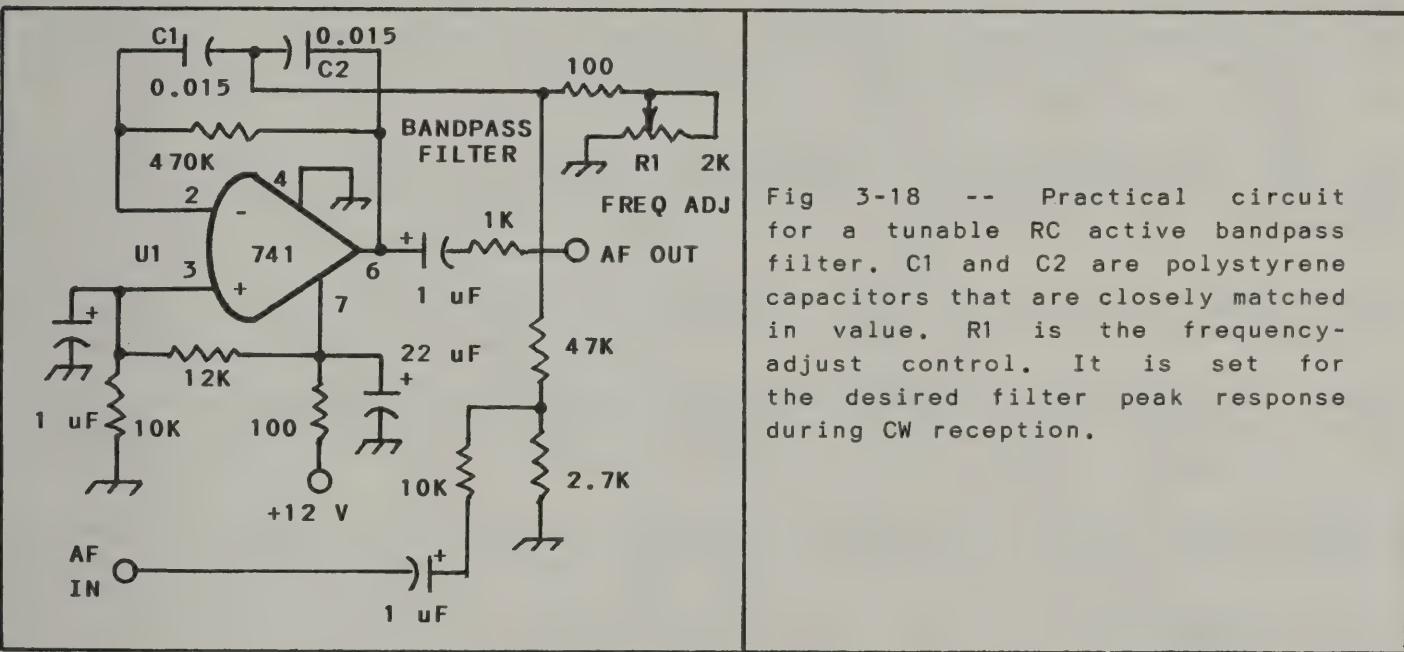


Fig 3-18 -- Circuit A is for a single-pole low-pass RC active filter that has a frequency cutoff at 700 Hz for CW reception. R1 and R2, as well as C1 and C2 must be closely matched in value. Circuit B is for an adjustable audio notch filter. R3 is used to select the notch frequency. R1 and R2, along with C2 and C3 need to be matched closely in value. Normally, a 5% variation in values is satisfactory.

A practical example of a 2-pole RC active audio bandpass filter is given in Fig 3-9. The low-pass filter of Fig 3-18A is useful for CW reception if the cutoff frequency is set for 500-700 Hz. For SSB operation the cutoff frequency should be 1800-2000 Hz. The low-pass response helps to attenuate high-pitched QRM, and it removes audio hiss noise at the output of the receiver. Both filters in Fig 3-18 require quality capacitors for C1 through C3. Polystyrene or mylar capacitors are suitable for preventing Q degradation and drift. The component values specified in Fig 3-18B provide a notch range that is suitable for CW reception. Both Fig 3-18 filters work best when they are located directly after the receiver audio gain control or first audio amplifier. They should not be subjected to high AF signal levels if they are to perform in an optimum manner.

A tunable RC active bandpass filter is illustrated in Fig 3-19. The same component-selection rules given for the Fig 3-18 circuits apply to this filter. This filter lends itself nicely to QRP receivers because of its simplicity and good performance during CW reception. R1 is adjusted for the desired peak audio frequency.



Since the circuit above has a bandpass type of response, it rejects frequencies above and below the frequency of interest. This filter is especially helpful when the receiver does not have a CW filter in the IF system.

Digital ICs

Digital ICs and their related circuits are beyond the intent of this book, since we are dealing primarily with RF and analog circuitry here. An excellent reference for digital ICs is **The ARRL Electronics Data Book**, second edition, 1988. Another excellent reference is **The Art of Electronics** by Horowitz and Hill, published by Cambridge University Press, Cambridge, MA. Chapter 8 of the book is devoted entirely to digital and logic circuits and their design.

Chapter Summary

I encourage you to experiment with the practical circuits outlined in this chapter. These topics are expanded in **Solid State Design for the Radio Amateur**.

Glossary of Chapter Terms

Active Circuit - A circuit that requires an applied operating voltage.

Active Filter - An audio filter that uses transistors or ICs.

AGC - Automatic gain control (also AVC, for automatic volume control). Used in receivers to maintain a nearly constant audio-output level. Derived from sampled IF or audio energy, then rectified and applied to one or more IF stages as a control voltage which varies with the incoming signal level.

Array, IC - An integrated circuit that contains two or more bipolar transistors to which access is available at the IC pins. Multipurpose ICs that may be connected for a variety of applications. A group of transistors on a common substrate.

Balanced Modulator or Mixer - A circuit that contains two or four diodes or transistors that has a balanced (push-pull configuration) input and output circuit. Provides greater isolation between the three ports than is possible with a single-ended balanced modulator or mixer.

BFO - Beat frequency oscillator. Operates at the receiver IF and is used to supply a carrier for SSB signals or a beat note for CW signals. Applied to the receiver detector.

BIFET, IC - An operational-amplifier IC that has JFETs at its input to ensure a high input impedance and low noise.

Bifilar, transformer - A transformer that has two identical windings which are wound on the core material at the same time, side by side or as a twisted pair of wires. A trifilar winding is the same, but with three wires.

Broadband Transformer - A transformer that is not tuned to a particular frequency. Wound in such a manner as to ensure efficient performance over several octaves of frequency, such as from 1.8 to 30 MHz. Generally wound on a ferrite core that has high permeability.

Conduction Voltage (diode) - Also called barrier voltage. The forward voltage value that causes current to flow through the diode junction. This occurs at 0.3-0.4 V for germanium diodes and 0.6-0.7 V for silicon diodes.

Conversion Gain - Relating to mixers. A measure of the output signal level with respect to the input signal level, and expressed in dB. A mixer may have conversion gain or loss, depending upon the devices used and the circuit design.

dBm - Power expressed in dB (decibels) as referenced to 1 mW. 0 dBm equals 1 mW.

DC Receiver - Direct-conversion (synchrodyne) receiver that uses a product detector instead of a mixer. Output from the detector is at audio frequency. The LO operates at the same frequency as the incoming signal, and it is tunable.

Diode Ring, mixer - A mixer or balanced modulator that uses four diodes in a doubly balanced circuit.

DR - Dynamic range. The measure of a receiver's performance in the presence of strong signal levels. Refers to the performance of the receiver input stages in particular (RF amplifier and mixer) in terms of IMD (intermodulation distortion generated within the mixer). The greater the DR the greater the signal-handling ability of the receiver.

Filter, audio - An LC (inductance-capacitance) passive filter or an RC (resistance-capacitance) active filter that is used to shape the audio-frequency response in a receiver. The filter response may be high pass, low pass or band pass, depending upon the requirements.

Filter, notch - An audio or RF filter that rejects a single frequency while passing the frequencies above and below the notch frequency. Used to reject a single tone, such as that caused by a carrier (heterodyne) during CW and SSB reception.

Gain - The measure, in dB, of a circuit's ability to amplify signal energy. May be expressed as AV (voltage gain) or power gain.

Gain, overall - The net gain in dB of a composite circuit, such as that measured from the input of a receiver to the audio-output port.

Heat Sink - A heat-conductive device (usually aluminum) to which a solid-state device is affixed to conduct heat away from the device and lower its junction temperature. May also be a small cooling device that is attached to the semiconductor, such as a press-on heat sink.

IC - Integrated circuit. A multi-pin semiconductor that has many transistors, diodes and resistors on a single silicon wafer.

IF - Intermediate frequency. Output frequency of a mixer. The sum or difference frequency of the signal frequency and the LO frequency.

LED - Light-emitting diode. Conduction voltage is approximately 1.5. Available with lenses of various colors and is used as an status-indicating light or pilot light.

LO - Local oscillator. Generally relating to the tunable or fixed-frequency oscillator voltage that is applied to a mixer in a heterodyne type of circuit.

LSI - Large scale integration. Refers to large ICs with hundreds of transistors, diodes and resistors on a common silicon wafer.

MIXER - A device that mixes two frequencies to produce a third (IF) frequency. Used in receivers, converters and transmitters.

NF - Noise figure. A figure of merit, expressed in dB versus frequency, for an amplifier or mixer. The noise is generated within the device as a result of current flow and temperature. Also the measure of a receiver's performance in terms of internal noise versus the level of the incoming signal.

Phasing, transformer windings - The manner in which the windings are connected to a circuit with respect to the required polarity of the circuit, which may, for example, require a 180-degree phase difference for the applied signal.

PIN Diode - A special variable-resistance diode that is designed for use at VHF and higher to function as a switch.

Product Detector - Used for the detection of AM, SSB, SSTV and CW signals in a receiver. The output is the product of the two input signals (IF and BFO), and is at audio frequency.

Substrate - Foundation upon which the elements of an IC are formed. A wafer of silicon material that is the foundation of an IC.

Varactor or VVC Diode - A voltage-variable capacitance diode that is used as a frequency multiplier or tuning device in place of a mechanical tuning capacitor. The junction capacitance may be changed in accordance with a variable applied reverse voltage (positive voltage applied to the cathode of the diode).

VBFO - Variable beat-frequency oscillator. Tunable oscillator used in a DC receiver.

CONSTRUCTION PRACTICES

CHAPTER 4

It will be helpful if we cover a number of subjects that relate to component selection, component use and general workshop practices before we move ahead to the chapters that describe practical projects. Workshop routine is as important to you as is the matter of circuit design: A well-designed circuit can become a dismal failure if the project is constructed wrongly.

CHOOSING THE RIGHT COMPONENT

Some experimenters are confused about the type of capacitor, resistor, diode or some other component that goes into a circuit. This is especially significant when an inexperienced builder attempts to duplicate a circuit that is published in *QST*, *The ARRL Handbook*, or some other Amateur Radio technical publication. Perhaps, for example, you have never heard of a polystyrene capacitor. You may wonder if it is suitable to substitute a disc ceramic or silver-mica capacitor. Resistors can cause consternation also. The author or designer may specify the resistors as 1/4-W carbon composition units. You are unable to find this style of resistor listed in your catalogs. Rather, you find 1/4-W carbon-film resistors offered by the vendor. Can these resistors be interchanged in a circuit without impairment of performance? You may find also that the designer specified several 0.015-uF disc capacitors, but your parts collection does not contain capacitors of that value. Instead, you have a number of 0.01-uF capacitors. Can they be used as bypass capacitors? Yes, they can in 99% of the circuit applications. This chapter addresses questions that fall into the foregoing category.

Capacitor Selection

Some circuits require capacitors that are stable in terms of maintaining their specified values versus changes in internal and external operating temperature. Oscillators and filters, in particular, have this requirement. A single unstable capacitor in a VFO can totally spoil the performance by causing frequency drift. NPO ceramic capacitors are perhaps the best choice by amateurs for all stability-critical circuits. They maintain their values well when temperature changes occur. Heating occurs not only from temperature excursions outside the capacitor (ambient) but also from internal circulating current. RF energy that flows through a capacitor causes it to heat internally, and this can change its value too. Therefore, it is wise to use two or more capacitors to replace a single one (parallel capacitors that have a net value equal to the required single unit). This provides greater internal area for the circulating current, thereby minimizing the effects of heating. For example, a VFO may require a 100-pF fixed-value NPO capacitor in parallel with the VFO coil. If you use three 33-pF NPO in parallel at this circuit point you will have 99 pF of capacitance, but the circulating current now has three times the capacitor area to pass through. This may not seem to be cost-effective for a small project, but the rewards are well worth the extra

minor expense.

Polystyrene capacitors are quite stable also, and they are inexpensive. They work well in oscillators and filters up to, say, 14 MHz. They have a slight negative-drift characteristic. This is beneficial when the oscillator coil has a core that is made from powdered iron: Powdered iron has a positive drift characteristic, so the negative drift property of the polystyrene capacitor often compensates for the positive drift of the core material.

Silver-mica capacitors may be used in frequency-critical circuits, but they are unpredictable in terms of negative or positive drift with changes in temperature. You may find that hand-picked silver micas work well in your VFO, but it may be necessary for you to try a number of same-value capacitors in the circuit before you obtain a set that provides stability. The task can be a tedious one!

Disc ceramic capacitors are excellent for nonstringent coupling and bypassing assignments at RF within the MF and HF bands. Certainly, they are better to use than is the situation with tubular paper or mylar capacitors. The latter types have a certain amount of internal inductance, owing to the manner in which they are constructed. When unwanted series inductance is presented in a circuit it can nullify the function of the capacitor. In some instances it may be so severe that no effective bypassing takes place, especially at the higher frequencies where small unwanted inductances cause major problems. Tubular or rectangular paper or mylar capacitors are, however, excellent units to use in audio circuitry.

Capacitors that are used at VHF and above work best if they are the leadless, monolithic chip-capacitor type. These devices may be soldered directly to the PC-board conductors to ensure minimum unwanted inductance. The leads on disc or dog-bone ceramic capacitors introduce considerable inductance at VHF and higher. If you must use them in your VHF project, try to clip the leads as close to the body of the capacitor as is practicable. If you short-circuit the leads of a disc ceramic capacitor you can check it with a dip meter to find the frequency at which it is a resonant circuit. The lower this frequency the greater the stray inductance.

Tantalum versus Electrolytic

Circuit diagrams often indicate capacitors for which the polarity is marked. These may be electrolytic or tantalum types. For the most part these capacitors are interchangeable for low frequency and audio work. Tantalum capacitors have very little stray inductance as opposed to electrolytic types. This makes the tantalum capacitors the better choice for circuits where highly effective bypassing or coupling is required. It is vital that you always observe the polarity of the capacitor. An incorrectly installed electrolytic or tantalum capacitor may overheat or blow up! Make certain also to select a capacitor that has a voltage rating somewhat greater than the supply line value. For example, if your circuit operates at 12 volts, use a 16- or 25-V capacitor.

Variable Capacitors

Air (dielectric) variable capacitors are the best ones to use in VFOs. They are the least prone to temperature-caused drift. Ceramic and plastic trimmers (even

NPO types) are not known for stability in critical circuits. Even though they may be relatively immune to value changes from heating, they are not stable in a mechanical sense. Vibration can change the capacitance, and changes in ambient temperature may cause the movable part of the trimmer to shift. Glass piston trimmers are a better choice if you're willing to spend the extra money needed to buy them.

VVC (varactor) diodes are the least stable of the available capacitors that we may use in an oscillator. They undergo temperature change from external and internal heat excursions. The dc and RF current that flows through a VVC diode is the cause of internal heating.

The air variable capacitor that you use for tuning your VFO should be the double-bearing type (a rotor bearing at each end of the capacitor) in order to ensure mechanical stability and minimum stray inductance. It should rotate freely to facilitate easy, non-lumpy tuning. Avoid using air variables that have aluminum plates when building a tunable oscillator. Aluminum is affected substantially by changes in heat. Plated brass vanes are more temperature-stable than aluminum ones.

Resistors

We discussed earlier in this chapter the matter of carbon-composition resistors versus carbon-film resistors. Unfortunately, most manufacturers of resistors have opted for the newer carbon-film type of resistor. The older solid-carbon resistors are becoming as scarce as fur on a frog. The carbon-film resistor has a helix of carbon ribbon wound around the insulating form. The greater the resistance the more turns of resistive film required. Fortunately, these resistors do not exhibit significant inductance, and they are entirely adequate in most HF circuits. The internal inductance, plus that of the resistor leads, can present a problem at VHF and higher. Today's industrial technology calls for the use of monolithic chip resistors at VHF and beyond. They solder directly to the PC board in a like manner to chip capacitors. These modern resistors are not widely available as surplus, so we must make do for the present with the older style of resistor with pigtails. Keep the leads as short as practicable, and you should be able to obtain acceptable circuit performance with them. You can identify the carbon-film resistor by its structure. These units have slightly enlarged ends (heads), whereas the older carbon-composition resistors have bodies that are uniform in size.

Power resistors are available in NIT (noninductive type) and standard format. The latter style uses nichrome wire wound around the insulating form of the resistor. These are highly inductive, which makes them unsuitable in RF circuits. The NIT resistor should always be used for resistive RF loads and dummy antennas. Globar resistors are also noninductive, but the resistance changes with heating. This is not usually a problem if a high-wattage Globar resistor is used in a very low-power circuit. In other words, a 10-W Globar resistor will exhibit minor resistance change when 1 watt of RF power passes through it.

Potentiometers need to be chosen carefully also. Many of the low-cost imported brands have a very thin layer of carbon on the inner element. Short-term use can wear away the carbon and make the control scratchy or intermittent. In other than breadboard circuits it is prudent to spend additional money for the purchase

of an industrial grade control, such as the Allen-Bradley brand. These controls are rated at 2 watts, whereas most other controls are specified for 1/4 or 1/2 watt, thereby indicating a thin layer of carbon under the slider. Pay close attention to the power rating of the control you use. An underrated control may burn out quickly when the related equipment is turned on. Determine the voltage and current associated with the control when making a selection. This is seldom a consideration for audio-gain controls where small currents are involved, but if a control is used, for example, with a three-terminal voltage regulator IC to vary the output voltage, considerable current will flow through the control.

Choosing the Right Transistor

Some builders feel that they must use the exact transistor specified in an article. This may be an important aspect of circuit performance at VHF and above, respective to noise figure and gain. It is wise in this instance to use what the designer recommends, even though a comparable transistor of a different number and brand may be available. Experimenters who are experienced in transistor selection can often pick a substitute transistor that will provide identical circuit performance. I do not recommend this to beginners or inexperienced builders.

Things aren't quite so critical for circuits that operate in the HF spectrum and lower. Circuits for these frequencies are more forgiving than those that are earmarked for VHF and higher. A good rule of thumb when making a selection for an MF- or HF-band circuit is to choose a transistor that has an f_T (upper frequency rating) that is five times or greater the proposed operating frequency. This ensures that ample gain will be available at the chosen frequency. Remember that the f_T is a frequency at which the transistor gain is unity or 1. In other words, it no longer amplifies the signal applied to it. A phenomenon associated with bipolar transistors is that the gain, in theory, increases some 6 dB per octave as the frequency is lowered. By way of example, if a given transistor is rated at 13 dB of gain at 30 MHz, it could have a gain of 19 dB at 15 MHz and so on. Although theory says this can happen, in practice it does not. There is, however, an increase in device gain as the operating frequency is lowered. This leads to potential instability (self-oscillation) problems at the lower end of the operating range. This may be the reason your RF amplifier is stable at, say, 14 MHz, but the circuit oscillates wildly at 3.5 MHz. Broadband RF amplifiers that have shunt-feedback networks do not suffer from this form of frequency sensitivity. As the operating frequency is lowered the effective feedback increases. This automatically reduces the gain of the amplifier. Some amplifier circuits include also a gain-leveling network at the amplifier input. This may consist of an inductor and a capacitor in shunt with the amplifier input. At the higher frequencies the network is "invisible," but as the operating frequency is lowered the network causes a reduction (loss) in input-signal power. This enables the amplifier to provide relatively constant gain across its frequency range. In plainer terms, this network presents a high reactance at the upper end of the operating range, but exhibits lower and lower reactance as the frequency in MHz is decreased.

Try to avoid using UHF-rated transistors for HF applications. This will help you to avoid excessive gain and the attendant instability mentioned above. Severe self-oscillation may destroy a transistor instantly, so try make an effort to choose the right transistor for your circuit.

A good all-around small-signal transistor is the popular 2N2222 or 2N2222A. Some acceptable substitutes for applications from dc through the HF spectrum are the 2N3904, 2N4400, 2N4401 other devices with similar electrical characteristics. Most of these transistors are interchangeable in circuits that call for one or the other types listed here. These are NPN transistors. Be careful when making a substitution so that you do not end up with a PNP device, which requires a different voltage polarity if the associated circuit is not changed to accommodate the PNP unit. I recommend that you obtain a transistor substitution guide if you intend to do a lot of experimenting. These books list the electrical characteristics of most of the common transistors. This enables you to locate other transistors that have similar or identical characteristics.

RF Power Transistors

QRP operators often use the 2N2222A or 2N4401 for RF amplifier circuits. These transistors and their equivalents can produce up to 0.5 watt of output power at +12 volts. They may be used in parallel to obtain up to 2 watts of output power. A number of large audio power transistors work well as HF RF power amplifiers, and they are inexpensive surplus devices. One example is the Motorola MPS-U02 which sells for under \$1 as surplus. This transistor can provide up to 2 watts of RF power at +12 volts if its tab is attached to a small heat sink. It has an fT of 150 MHz. The MPS-U05 is a trifle huskier, and is a good one also for RF service. Most power transistors that are intended for Hi-Fi amplifier use have high fT ratings, and they are not costly.

In a like manner, many low-cost power FETs, such as the IRF511, work well as HF-band RF amplifiers up to 20 meters. The IRF511 is designed for switching rather than RF applications, but it can deliver up to 15 watts of RF power when operated at +24 volts. Power FETs are available for RF service, but they are very expensive.

Toroids and Slug Cores

It is essential that we use the right core material for a given circuit. Slug-tuned coils generally contain a powdered-iron core that is used to vary the coil inductance. Various types of core material are available, and each is rated for a specific frequency range versus coil Q. The wrong core can destroy the circuit Q and spoil the circuit performance. The same is true when we use toroid cores. A useful reference to core materials and their optimum frequencies of operation is the J. W. Miller Co. catalog (19070 Reyes Ave., Box 5825, Compton, CA 90224). The Amidon Assoc., Inc. catalog (12033 Otsego St., N. Hollywood, CA 91607) lists the core characteristics for powdered-iron and ferrite toroids and rods. The message here is to avoid using nondescript cores that are offered as surplus or at flea markets, unless the seller specifies the core mix. Some slug and toroid cores are color-coded. For example, red indicates that high Qs may be obtained from 1 through 10 MHz. Yellow cores provide high Qs from 1 through 50 MHz. The core permeability varies with the core mix also. The higher the specified operating frequency the lower the permeability, and the lower the permeability the greater the number of coil turns required for a given inductance.

The decision to use ferrite versus powdered-iron cores must be founded on the circuit application. Powdered iron has a much lower permeability for an equivalent ferrite-core size. But, powdered-iron cores are more permeability-stable than

are ferrite ones. Avoid using ferrite in VFOs and filters, where all components must be temperature-stable. Ferrite cores saturate more quickly for a given power level than is the case with powdered-iron material. When saturation occurs the core has reached its maximum power-handling capability (flux density). The core becomes hot, and in the case of ferrite may fracture. Ferrite cores undergo a marked change in permeability when saturated, and the permeability may never return to its original value after cooling takes place. Powdered iron, on the other hand, does not undergo a major change in permeability when saturated. It does return to its normal value when it cools.

The advantage of ferrite versus powdered iron, for a given core size, is that higher permeability exists. This means fewer coil turns, and under some circumstances -- lower loss.

Ferrite is the core of choice for broadband transformers, including baluns. This is because the ferrite reduces the number of transformer wire turns, along with becoming "invisible" at the upper end of the transformer frequency range. At the lower end of the range, where higher effective inductance is needed, the core is "seen" by the winding. It is this phenomenon that allows the transformer to operate in a broadband manner.

An ideal VFO has no core material inside the VFO tuned-circuit inductor. However, yellow no. 6 powdered iron is the best choice if a toroid or slug is used. It is a highly stable core material, and provides a high tuned-circuit Q up to 50 MHz. An example of a good VFO toroid core is the Amidon T68-6 unit. The winding should consist of the largest wire size practicable and it should be cemented in place by doping it with two or three coatings of General Cement Polystyrene Q Dope or an equivalent high-Q glue.

How to Wind Toroids

A major area of confusion among inexperienced builders is how to wind a toroid inductor or transformer. People have asked me such questions as, "Is the winding wrapped around the outer perimeter of the toroid, or do you loop the turns through the core?" The answer is, of course, that the turns are passed through the core and wound around it progressively. Although the turns could be wrapped around the outside of the toroid (a difficult task, indeed!), the inherent self-shielding property of the toroid inductor would be lost. Other questions asked are "How do I wind the secondary with respect to the primary?" and "Should the winding occupy all of the core area or should it be bunched over a small part of the core?" Fig 4-1 shows some illustrations that relate to these questions.

Broadband transformers that are designed for use from 1 to 30 MHz generally contain a high permeability core material (800-900 μ). No. 43 core material from Amidon Associates fits this description. Fig 4-1 also shows a method you may adopt to fashion a high-power ferrite binocular core from ferrite toroids. This type of core is used generally for the output transformers in high-power solid-state RF amplifiers. One-piece binocular or balun cores are also available from Amidon.

Most ferrite toroids have sharp edges along the inner and outer perimeters. Care must be taken during the winding process to avoid cutting through the enamel insulation of the magnet wire. If you have a rock tumbler available, use it to

smooth the rough edges before you apply the winding. Alternatively, you may dip the core in clear polyurethane varnish before you use it. Large cores may be wrapped with insulating tape before the winding is added. These measures help to prevent damage to the wire insulation.

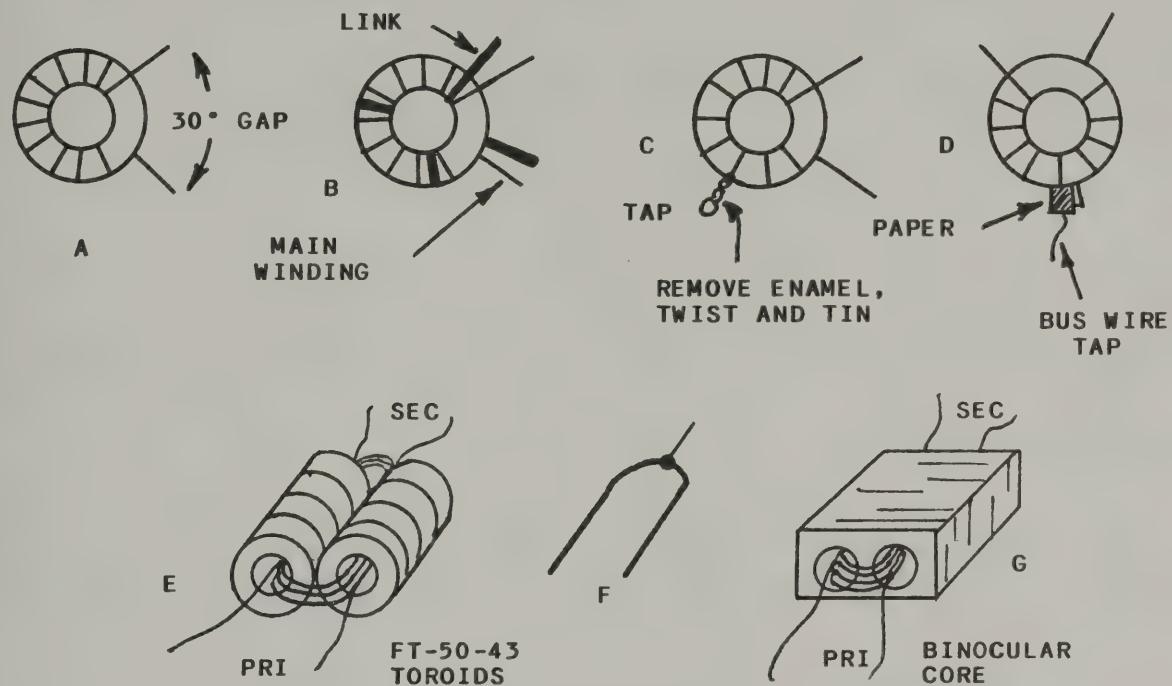


Fig 4-1 -- A 30-degree gap should exist between the ends of the winding (A). At B the link is shown as occupying the same core area as the main winding. Illustration C shows how a tap may be placed on a toroid winding. One turn of the winding is pulled away from the core, its insulation scraped away, and then it is twisted to form a connection point. The tap method at C requires the removal of the wire insulation on one coil turn. A piece of paper or other insulating material is slipped under the turn and a tap wire is soldered to the turn. Two rows of toroid cores may be cemented together with epoxy cement (E) to form a high-power RF transformer. At F is a U-shaped piece of insulated wire that may be used with the cores at E and G as the low-impedance primary of the transformer. One turn in high-permeability core material provides ample inductance at HF. The U-shaped wire represents one turn on the transformer.

Although Fig 4-1B shows the link wound over all of the primary or main winding, you can bunch the link turns at one end of the primary. It is generally held that the transformer efficiency is better when the windings are applied as shown at C of Fig 4-1. For tuned, narrow-band toroidal circuits I prefer to bunch the link turns at the cold or grounded end of the main winding. This reduces unwanted capacitive coupling between the two windings. This can be especially important toward reducing the transfer of harmonic currents.

Multifilar Windings

Bifilar, trifilar and quadrifilar transformer windings are commonplace in our modern solid-state circuits. Bifilar and trifilar windings are used in balun transformers for matching amplifier input and output impedances to the amplifier loads. They are used also for antenna baluns. Bifilar indicates a two-wire winding, trifilar signifies three windings and quadrifilar transformers have four windings. These windings are placed on the transformer core at the same time. They may be laid on the core in parallel with one another, or they are sometimes twisted together (about eight twists per inch) before the winding is placed on the core. The twisting of the wires is done easily with a hand-operated drill or an electric one that is slowed to a few RPM. One end of the wire bundle is placed in a vise and the other end is clasped in the chuck of the drill to accomplish this.

Identification of the multifilar windings may be carried out by "ringing" them out with an ohmmeter. If you use a different color enamel wire for each of the windings the job will be easier. The magnet wires may each be spray painted a different color before they are wound on the core. This makes it an easy matter to tell which wire is which when connecting the transformer to the circuit.

Checking Toroid Inductance

The inductance of a toroid- or rod-core inductor can be determined beforehand if you know the AL factor of the core. This information is available from the manufacturer or vendor. Once this factor is known the correct number of turns can be obtained from

$$N = 100 \sqrt{L_{\mu H} \div A_L} \quad \text{Eq 4-1}$$

or

$$N = 1000 \sqrt{L_{mH} \div A_L} \quad \text{Eq 4-2}$$

where N is the number of turns and AL is the manufacturer's turns versus permeability factor. Eq 4-1 is used generally for small inductances on powdered-iron cores, whereas Eq 4-2 relates to ferrite cores with their high permeability.

If you do not know the AL of particular core, you may learn what it is by placing an arbitrary number of turns on the core, then measuring the inductance. Once you know the inductance of the winding and the number of turns required for this inductance you can obtain the AL from

$$A_L = \frac{L_{\mu H} \times 10^4}{N^2} \quad \text{Eq 4-3}$$

You can measure the inductance of a toroid or other type of winding if you own a dip meter. This test is done by placing a known-value capacitor in parallel with the winding of unknown inductance. Wind a two-turn link over the unknown coil winding, then attach an identical link to this one as shown in Fig 4-2. The dip-meter probe is inserted into the external link, as shown. The dipper

is tuned for a dip in meter reading, which indicates tuned-circuit resonance. The XC of C (known) in Fig 4-2 is obtained from $XC = 1/(2\pi fC)$, where XC is the capacitive reactance in ohms, f is the frequency in MHz and C is in μF . Once you have this information you can learn the L_1 inductance from $L(\mu\text{H}) = XL/(2\pi f)$. At circuit resonance the XL is equal to XC .

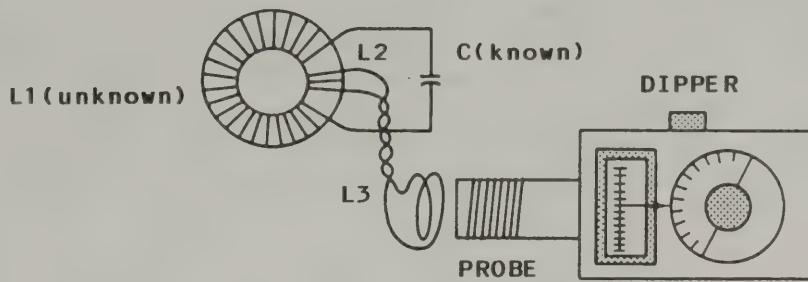


Fig 4-2 -- Illustration of how to use a dip meter to check the inductance of an unknown toroidal inductor. C is a capacitor of known value. L_2 is a temporary two-turn link and L_3 also has two turns. Circuit resonance is found with the dip meter. $XC = XL$ at resonance (see above text).

The test method shown in Fig 4-2 is adequate for most amateur work. You should be aware that the presence of L_2 and L_3 has some effect on the true inductance value of L_1 . In other words, the measured value will not be exactly the same as the coil inductance without L_2 and L_3 connected. The difference is rather minor at HF, and should not be a matter of great concern. The Fig 4-2 technique is necessary because of the self-shielding property of a toroidal coil. The link method of sampling may be avoided if the pigtails on C of Fig 4-2 are sufficiently long to permit inserting the dipper probe between the leads. The pigtails and the capacitor form a one-turn loop that can be probed with the dipper. The coil-inductance reading is more accurate if this is done.

Wire Size versus Toroid Turns

The Amidon Assoc., Inc. catalog has tables that list popular toroid-core sizes versus the maximum number of wire turns. The larger the wire the fewer turns possible on a given core. Also, there becomes a practical limit to the wire gauge, since large-diameter magnet wire can't be wrapped on a small core without ending up with turns of unequal diameter. This is caused by the stiffness of the larger wire. Furthermore, large-diameter wire requires considerable tension when winding it on a toroid core. The excessive stress can cause a core to break.

Wire Length Before Winding

You may wonder how much wire to cut from the roll for, say, a 20-turn toroid winding. Wrap one turn of wire on the core, then remove it and measure the length. Multiply this by the number of turns, then allow four additional inches of wire to ensure adequate lead length at the ends of the winding. Be careful to avoid kinks or loops in the wire as you are winding the toroid. If they develop, be sure to straighten the wire before proceeding with the winding.

Flux Density of Cores

Magnetic cores are rated in gauss or flux density. This is the density of the lines of force, as measured at a cross section of their flow. This may seem rather complicated to ponder, but it is an important aspect of any design that concerns toroids or other magnetic-core devices. This parameter is specified as B . $B(\text{max})$ is the maximum core flux density as expressed in gauss. Cores are rated in gauss, and this data may be found on the manufacturers' specification sheets. We must use care to not exceed, in an operating circuit, the rated flux density of the core. Exceeding this limit causes core saturation and heating. During saturation a core can produce harmonics and its permeability will change. Best performance occurs when we allow a substantial margin between the maximum gauss of the core and the flux density, $B(\text{max})$ developed in our circuit. If you know certain factors that relate to the core and circuit you are using it is possible to calculate the $B(\text{max})$ from this equation:

$$B(\text{max}) = \frac{E_{\text{rms}} \times 10^8}{4.44fNA_e} \text{ gauss} \quad \text{Eq 4-4}$$

where $E(\text{rms})$ is the applied ac or RF voltage, A_e is the equivalent area of the magnetic path in cm^2 , f is the frequency in Hz and N is the number of inductor turns. A good rule of thumb is to choose a core that is larger than that indicated by Eq 4-4. If the core you are using runs warm or hot, replace it with a larger core. A core should never be more than moderately warm to the touch after five minutes of operation. Considerably greater information on this subject and the general theory and application of magnetic cores may be found in D. DeMaw, **Ferromagnetic Core Design & Application Handbook**, 1981, Prentice-Hall, Inc., Englewood Cliffs, NJ 07632. Available also from Amidon Assoc., Inc., N. Hollywood, CA 91607.

Table 4-1

Core Material	Permeability	Freq. Range	Color
0	1	100 to 300 MHz	Tan
1	20	0.5 to 5.0 MHz	Blue
2	10	2.0 to 30 MHz	Red
3	35	0.05 to 0.5 MHz	Gray
6	8	10 to 50 MHz	Yellow
10	6	30 to 100 MHz	Black
12	4	50 to 200 MHz	Grn & Wh
15	25	0.1 to 2.0 MHz	Red & Wh
Data for the most popular Amidon Assoc. and Micrometals Corp. powdered-iron toroid cores. Each core type may be used below the lowest indicated frequency to obtain high Q. The upper frequency limits are those where the inductor Q starts to drop.			

Table 4-2

Number of turns for a single-layer toroid winding

Wire Gauge	20	22	24	26	28	30	32	34	36
Core OD (inches)									
0.12	1	2	4	5	8	11	15	21	29
0.16	3	3	5	8	11	16	21	29	38
0.20	4	5	6	9	14	18	25	33	43
0.25	5	7	11	15	21	28	37	48	62
0.37	12	17	23	31	41	53	67	87	110
0.50	21	28	37	49	63	81	103	131	166
0.68	28	36	47	61	79	101	127	162	205
0.80	39	51	66	84	108	137	172	219	276
1.06	45	58	75	96	123	156	195	248	313
1.30	66	83	107	137	173	220	275	348	439
2.00	109	139	176	223	282	357	445	562	707
2.25	123	156	198	250	317	400	499	631	793

Winding information for the most popular toroid cores. The number of turns indicated above is for completely filling the core. In practice a 30° gap is provided between the ends of the winding, and this must be taken into account when using this table. Manufacturers' assigned core numbers relate to the OD listed above. A T50-2 powdered-iron toroid has a 0.5-inch OD. The number 2 signifies the core material. An FT-50-43 ferrite toroid has an OD of 0.5 inch. The 43 lets us know the core mix.

PC BOARDS and BREADBOARDS

Those of you who are new to experimenting may be afraid to attempt the layout and etching of a PC board. Perhaps the most important consideration is that the board will do the job intended. It need not look like a work of art. After all, it will become buried in a cabinet when the circuit is assembled on it. Chances are that you will be the only person to see your Ugly Duckling, should it take on an unprofessional appearance. A little experience with making your own boards will lead to a quality product, so don't avoid this satisfying part of your workshop experience.

Doing the Layout

Collect the components that mount on your circuit board. Use lined quadrille paper when drawing your PC-board pattern. Sketch the layout by placing each part

on the quadrille paper, stage by stage, to learn the spacing you need between copper pads. Draw an outline for each pad as you progress, and then draw in the connecting copper foils that join the various isolated pads. Be careful to avoid having components cross over one another. Resistors and some of the other components can be mounted vertically on a PC board to save space.

When you have completed your sketch of the board layout set it aside and cut a piece of PC board to the size of your pattern. Next, cover the copper side of the board with masking tape. Press the tape firmly in place by rolling a pen back and forth over the tape surface. It must adhere firmly to the copper. Now, lay a piece of carbon paper (carbon side down) on the masking tape. Put your sketch atop the carbon paper and anchor it in place with two strips of Scotch Brand tape. Trace the pattern with a ball-point pen, pressing down firmly as you trace. Remove the sketch and observe the pattern that you have transferred to the masking tape. Be sure that all of your pattern has been transferred.

Your next job is to cut away the masking tape, using a hobby knife, for those areas that are to be etched from the copper surface of the board. Place the PC board in etchant, agitating the work every three minutes, until all of the unwanted copper is gone. Wash thoroughly and remove the remaining masking tape. Polish the copper with steel wool. Your board is now ready to drill. Use wide masking tape when you cover the board copper. Seams caused by using strips of tape side by side (overlapping slightly) do not seal well against the etchant solution. I have used ducting tape successfully as an etch-resist material. Various kinds of adhesive-backed materials lend themselves to the foregoing application. Don't be afraid to experiment.

Perf Board Construction

If you do not wish to use etching chemicals you may use Perf Board (phenolic board material that has many holes, but no copper) as the base for your circuit. In fact, you can make your own perforated boards from Formica, linoleum tile and other durable material that has insulating properties. Drill rows of holes in the material to facilitate mounting the components. The hole size should be 1/16 inch in diameter.

You can now mount your components by passing their leads through the holes in the board. Bus wire may be used to join the related components, and to form ground and voltage bus conductors.

Piggy-Back Islands

A circuit board can be made easily by cutting many tiny squares of single-sided copper-clad PC board stock. These little squares become isolated pads or islands when they are glued to a blank piece of PC-board stock. Quick-setting epoxy cement works well for this application. The components can now be soldered to the various isolated pads. Use no. 22 or 24 tinned bus wire to join the circuit-related pads.

Grid of Squares Boards

A blank piece of copper-clad PC board may be cut with a hacksaw to develop a grid of square or rectangular isolated pads. Place the work on a flat surface and cut across the copper side of the board with a sharp hacksaw blade. Turn

the work 90 degrees and cut across the board several times again. Be sure to cut all of the way through the copper to ensure that your islands are free and clear of the adjacent ones. Solder the components to the pads and join the appropriate ones with bus wire.

The Etching Process

Ferric chloride is the most popular etchant for amateurs. It is a brown fluid that leaves stains, so handle with care. **Warning:** Avoid contact with the skin, eyes or mouth. Wear rubber gloves when working with any chemical. Another etchant is ammonium persulphate. This chemical produces a clear solution that does not stain objects that come in contact with it.

You can purchase ferric chloride and ammonium persulphate crystals from various amateur outlets. Exercise extreme care when mixing them with water. Follow closely the instructions. Ferric chloride crystals, in particular, must be added to the water slowly. If not, the mixture becomes hot and may spurt out of the container with a "whoosh!" If this occurs your face and hands can become spattered with a dangerous chemical. Premixed ferric chloride is available at Radio Shack stores and other outlets. It is entirely stable, once it has been mixed.

The circuit board to be etched should be cleaned thoroughly before immersing it in the solution. Steel wool works nicely for cleaning the copper. The etchant should be heated to 90-110 degrees F to assure fast action. Typical etching time is 15 to 30 minutes at this temperature, depending upon the thickness of the copper and the age of the etchant.

It is important to agitate the work every few minutes while the etching process is taking place. This removes the deposit of exhausted copper and allows the etchant to reach the copper surface of the board. An aerator for a fish tank works well to keep the solution agitated. The etchant may be brought to the desired temperature by placing the container on a warming plate of the type used to keep food warm.

When the etching is completed, wash the board with hot water and dry it with a towel. It is now ready to use, unless you want to tin plate it. Tin-plating powder is available from some amateur radio outlets. You simply mix it with a prescribed amount of water, stir in the chemical, and store the solution in a plastic container. To add tin plating to your PC board, simply lay the board in the solution for approximately 15 minutes at room temperature. The copper surface must be clean and free of finger marks and oil if you are to obtain a flawless plating job. Rinse the board in hot water and dry with a towel. The addition of tin plating prevents the PC board foils from tarnishing, and it makes soldering easier: Solder adheres to plated boards more quickly than to bare copper. **Warning:** Exhausted etchants should never be poured on the ground or down the sink drain. Keep them in bottles and dispose of them through your area waste-disposal facility. Identify the chemicals as "hazardous waste."

"Universal" Breadboards

You can create your own pattern for a "universal" breadboard. I keep several of these on hand for prototyping and finished projects. They consist of a PC board with numerous isolated pads. Some of them contain sites for 8-pin ICs as

well, while others have provisions for 16-pin ICs. A variety of these useful breadboards is available from FAR Circuits (N9ATW), 18N640 Field Court, Dundee, IL 60118 at this time. A brochure and price list is available from the supplier.

Certainly, a breadboard of this type eliminates the need to design a PC-board pattern for each new project. This assumes that the project in question is relatively simple. Complex circuits, such as high-performance receivers, deserve a more elegant and elaborate PC board. Fig 4-3 provides an etching pattern for one of the breadboards that I use frequently.

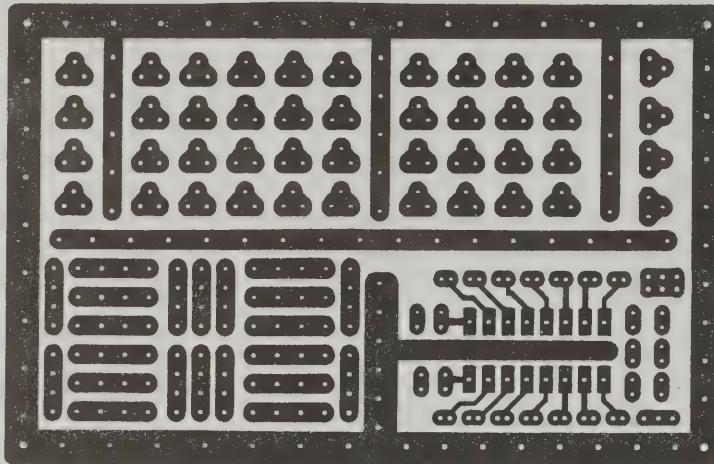


Fig 4-3 -- Scale etching pattern for a universal breadboard. This W1FB design includes a + voltage bus at the center of the pattern. Various ground foils come from the outer perimeter of the board to permit short leads for grounded components. A 16-pin DIP IC site is at the lower right. No. 60 holes are drilled in the copper pads to facilitate parts installation.

I have built a number of test-equipment items, QRP transmitters and some simple receivers on boards with the Fig 4-3 pattern. If you wish to have a ground plane on the component side of this breadboard you can use double-sided (copper on both sides) PC-board stock. A small drill may be used to remove the copper from around the holes (ground-plane side) that accommodate leads that are not grounded.

TEC-200 Etch-Resist Transfer Film

The Meadowlake Corp., 25 Blanchard Drive, Northport, NY 11768, sells sheets of 8-1/2 X 10 inch clear film that is used to photocopy patterns such as that in Fig 4-3. A dry-copier or any-paper copier is used for this. The PC pattern appears on the clear film, and this image is transferred to a blank piece of circuit board by using a household iron. In other words, this is a heat-transfer method. The black image then appears on the PC board and serves as the etch resist agent. The only gimmick associated with this quick-and-easy process is that the artwork that is placed in the copy machine must be a mirror image of the actual PC-board pattern desired. I use TEC-200 frequently for one-shot circuit boards. Instructions are supplied with the TEC-200 film.

I have attempted to use overhead-projector clear-film sheets for the above process. I was unable to transfer the pattern with an iron, although I have heard that this is possible. You may want to experiment with this type of film.

Home-Made Heat Sinks

Commercially made extruded heat sinks are costly. Worse still, it is not always easy to find a heat sink of the correct area for a particular project you are developing. Amateurs have always been known for their ability to innovate, and being able to create home-made heat sinks fits that description. Many ordinary materials are at hand for this frequent necessity. For example, hardware store aluminum angle stock works well when cut to an appropriate length. It is easy to bolt to a PC board, and it may be drilled for a press-fit to accommodate the body of, for example, a TO-5 or TO-39 transistor. Devices of the TO-220 class (with the metal mounting tab) may be bolted directly to this type of heat sink. Fig 4-4 shows various methods for creating home-made heat sinks.

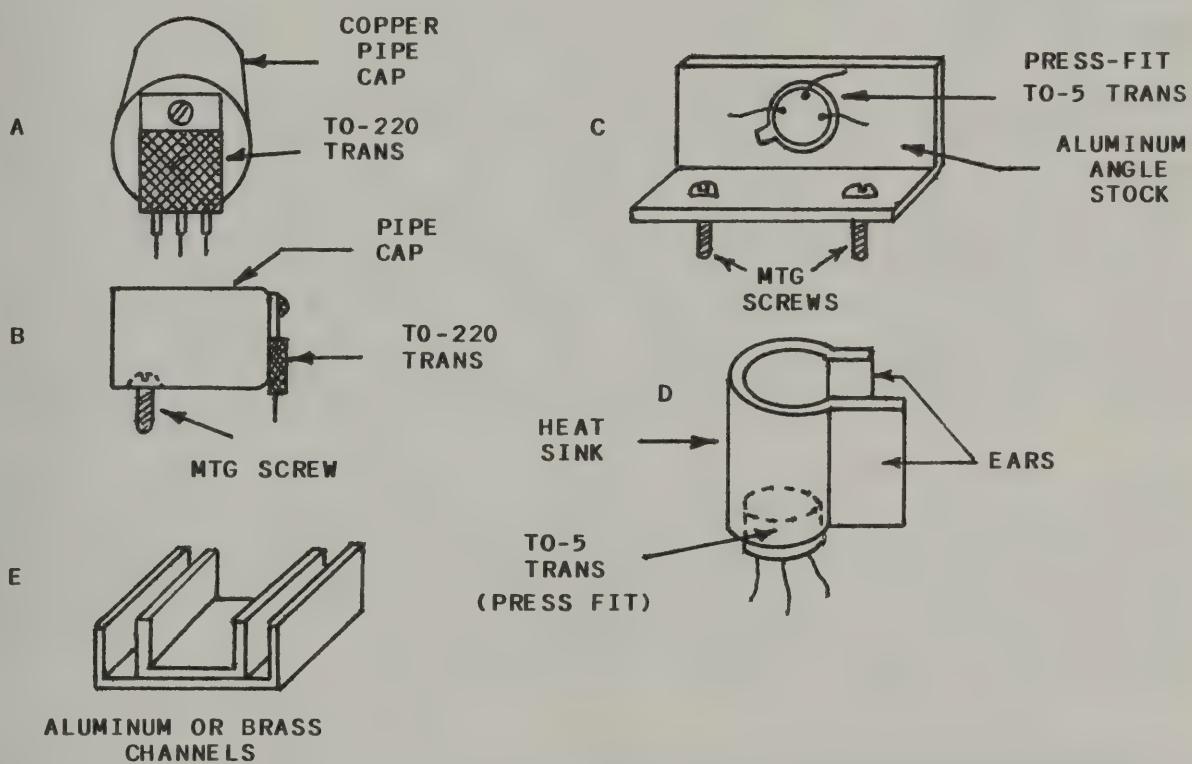


Fig 4-4 -- Examples of home-made heat sinks that are fashioned from readily available materials. A 3/4-inch copper pipe cap is shown at A and B. Hardware store extruded aluminum angle stock may be used, as shown at C. The hole for the TO-5 or TO-39 transistor is made slightly smaller than the transistor body to allow a press fit. Illustration D shows an aluminum or brass heat sink that is shaped around a drill bit that is slightly smaller in diameter than the transistor body. The ears may be made large to increase the heat-sink area. Two (or more) aluminum channels may be made from 16-gauge brass or aluminum stock and bolted together to form a large-area heat sink for TO-3, TO-220 and other large transistors. Silicon grease should be used between the channel surfaces.

When using heat sinks it is important that you apply a thin layer of silicone heat-sink compound to the mating surfaces of the transistor and heat sink. This helps to ensure efficient heat transfer from the transistor to the sink. The transistor needs to mate firmly with the heat sink. Do not use excessive torque when affixing any transistor to its heat sink. A snug fit will suffice.

Small transistors with plastic bodies, such as those in TO-92 cases, need heat sinks in certain critical applications. A good example is when we use one or more 2N2222 transistors as QRP RF power amplifiers. It is not easy to attach a heat sink to the half-round body of such a transistor. One method is to use an expended 22-caliber rifle shell as a heat sink. It can be compressed slightly at the open end to provide a snug fit on the transistor body. Quick-setting epoxy cement is then allowed to flow between the transistor and the shell casing. One or two drops is ample. This affixes the transistor to the heat sink and helps to assure the transfer of heat, although epoxy is not an especially good conductor of heat. Brass or copper tubing may be substituted for shell casings. TO-92 plastic transistors may simply be epoxy cemented to a small tab of copper, brass or aluminum when a heat sink is required. Clamp the transistor to the metal until the cement has hardened.

HOME-MADE COIL FORMS

Modern shotgun shells with their hard-plastic sleeves make excellent coil forms. Remove the discharged center-fire cap from the expended shell by driving it from the inside out with a metal punch and hammer. The resulting hole is now available as a mounting point when using a 6-32 screw and nut. Small holes can be drilled in the plastic sleeve to anchor the ends of the winding. I place two small holes side by side at each end of the winding area. I thread each end of the winding through the two holes. This keeps the turns in place and allows the ends of the winding to project outward from the coil form. Various gauges from 410 to 12 gauge are suitable. **Warning:** Do not attempt to remove the contents or cap from a live shell! Use only casings that have been discharged in an appropriate weapon.

Plug-in coil forms can be made from expended shotgun shells. After removing the discharged cap, enlarge the hole to accommodate an RCA phono plug. Solder the plug to the brass base of the shell with the tip of the phono plug pointing outward from the shell. The brass base and phono plug outer body are made common to one end of the coil. The remaining coil wire is soldered into the prong of the phono plug. The smaller shells (410 and 20 gauge) are best for this application because of their smaller size and reduced weight.

PL-55 phone plugs with plastic outer covers are also suitable for use as plug-in coil forms. The coil is wound on the plastic end cap and glued in place with polystyrene Q Dope or two coatings of polyurethane varnish. A stereo plug allows you to have three coil connections, whereas a mono type of plug provides only two coil terminals.

What About PVC Pipe?

PVC pipe is often acceptable for use as coil-form stock. I do not recommend it where high RF voltages are present, since it can melt or burn at high RF power levels. An example of a questionable application is that of a top-loading coil for a vertical antenna. I melted a 3-inch diameter PVC loading-coil form in three minutes when using it for a loading-coil form at the top of my 160-meter vertical antenna. I was using 300 watts of RF output power. Nylon is a similarly poor for use where high RF voltages are present, and it should not be used for any VHF and higher circuit.

PVC pipe is okay to use for coil forms in low-impedance circuits and for QRP power levels at high-impedance points in an MF or HF circuit. I prefer the type of PVC

pipe that is specified for hot-water lines. It is harder and seems to have better insulating properties than does the softer cold-water pipe.

Insulating materials may be tested for dielectric quality by placing a sample of the material in a microwave oven and operating the oven at full power for two minutes. If the insulating compound does not melt, burn or become excessively hot, it should be satisfactory for most RF circuits.

Good Insulators

A number of modern plastics work well as insulators at RF and for circuits that contain high dc voltages. Plastics such as Delrin, polyethylene, Lexan and polystyrene are excellent. I recommend high-impact polystyrene tubing and rod for outdoor applications. This is a milk-colored plastic, whereas regular polystyrene material is clear. The former type does not shatter when placed under stress in cold weather. Your local commercial plastics dealer should have the aforementioned materials in stock. You may be able to purchase end lots and scrap pieces at low cost.

Glazed porcelain and steatite has long been an excellent insulating material at RF. Glass is also good. Unglazed porcelain and steatite accumulates moisture and dirt quickly, and this renders the material lossy and conductive.

Glass-epoxy resins (fiber glass) make good insulators also. I have used this material successfully for antenna insulators and coil forms. This substance is used as the base for most circuit boards of quality.

I do not recommend phenolic rod and tubing for outdoor applications because it is porous and absorbs moisture. It may be used if the phenolic coil form or antenna insulator is coated with glyptol, spar varnish or some other weather-protective substance. If you must use phenolic insulators I recommend that you do not utilize paper-base phenolic. Cloth-base phenolic is more durable and is less porous.

HOME-MADE PROJECT BOXES and CHASSIS

There is probably no more frustrating an undertaking than trying to find a small cabinet or chassis for your new project. The cost for commercial project boxes is fast becoming prohibitive for those of us who build a lot of gear. Also, it is not always easy to find a commercial cabinet that has the right dimensions for the project at hand. What, then, is the solution to this knotty problem? We can make our own foundations and boxes for pennies. We can size them to fit the need.

Food Containers as Cabinets

The next time you stroll through a food store take a few extra moments to examine the metal containers for the various edibles. It will become apparent that you may have overlooked an important hobbyist resource for many years! Sardines and other types of sea food are packaged in a variety of containers. Some are round, while others are rectangular. These serve nicely as small chassis for home-made test equipment, QRP rigs and power supplies. Fig 4-5 shows how this may be done. If you consume the contents of the container, you may consider the chassis a free commodity. If you don't wish to get involved with a food that doesn't appeal to you, move along to the cook-ware department. Here you will see all manner of small and large bread pans and cake tins. These items serve well as foundations for

home-built equipment. The finished product may be sprayed your favorite color of paint, or you may elect to cover the chassis with contact paper of your choice. Don't overlook the aluminum cookie sheets in the cook-ware department. These are relatively inexpensive, and they provide a source of aluminum for making cabinets and chassis. In a lighter vein, I don't recommend that you use Life Savers or bagels as toroid cores! But, getting back to metal food containers, one of the more popular QST projects of the 1970s was the "Tuna-Tin Two," which was a 1/4-W QRP transmitter for 40 meters: It was built on a tunafish can. Another popular project of that era was the W1VD "Sardine Sender," which was an 80-meter QRP transmitter assembled on a rectangular sardine can.

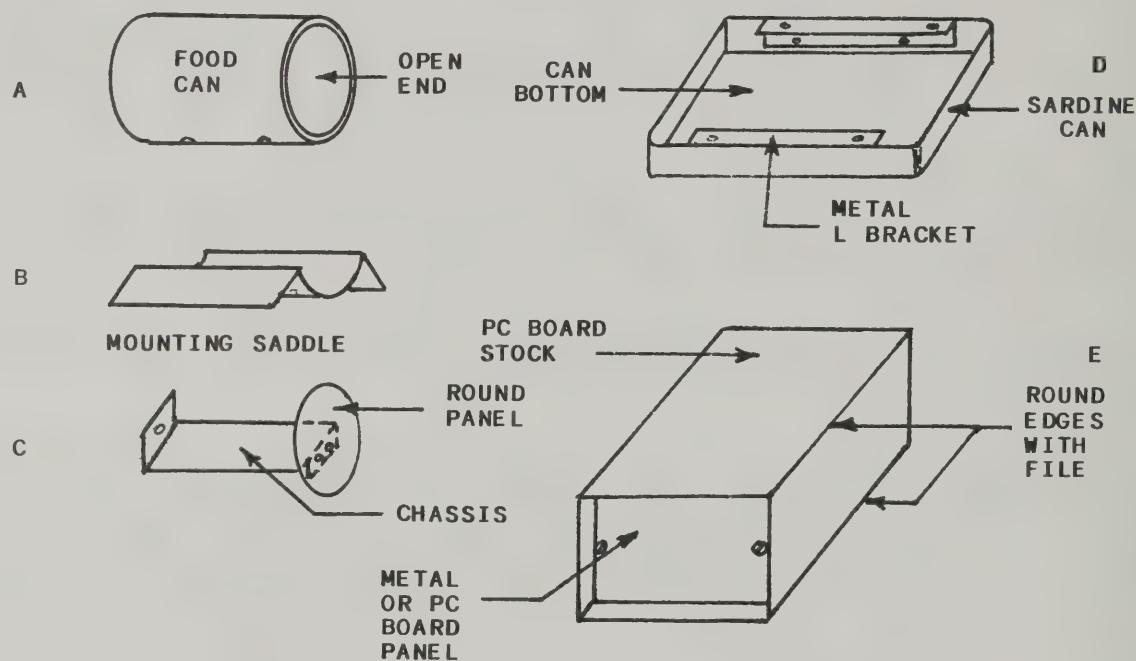


Fig 4-5 -- Simple, inexpensive chassis and boxes can be made by following these methods. A, B and C show how to use a food can as a project case. A mounting saddle attaches to the can with two no. 6 sheet-metal screws to keep the cabinet from rolling. A circular PC-board or metal panel (C) is attached to a PC board or aluminum chassis. The chassis is attached to the closed end of the can with a sheet-metal screw. At D is a sardine tin that is modified for use as a chassis. The PC-board project mounts at the open end of the can by means of two metal L brackets. Control holes may be drilled in the sides of the can. The box at E is made from sections of single- or double-sided PC board. The walls are joined by running a seam of solder along the mating edges. L brackets at the open end of the box permit affixing the panel.

Don't overlook metal recipe-card boxes when searching for a small cabinet. Check office-supply stores for other metal boxes that can be used as cabinets. Bond boxes with handles are excellent for housing portable amateur stations or test equipment that must be carried afield. I once knew a ham who built his entire ham station in a two-drawer metal file cabinet. Don't be afraid to break away from conventional trends when housing your circuits.

Surface Preparation & Finishing

Most of us try to impart the "commercial" look to our finished products. This facet of workshop practice can be as challenging and satisfying as is the design and debugging of our pet circuits. Obtaining a smooth and blemish-free paint job is no simple matter if you lack experience. Task no. 1 is getting the paint to adhere well to the raw surface. Some tedium and patience is necessary if we are to have repeated success with our paint jobs. The following steps are the ones I follow.

- 1) Scrub the raw metal or PC-board material with hot water, kitchen cleanser and a brush. Rinse it thoroughly in warm water, then dry it with a clean cloth. Do not touch the surface of the work with your fingers, lest oil from your skin create paint-resistant spots.
- 2) Paint will adhere much better if the surface of the work is abraded with a fine grain sandpaper before you clean the work. Move the sandpaper back and forth in a constant direction (not circular rotation) by pressing down lightly on the sandpaper block. Many fine grooves will appear as you work. These help the paint to stick to the work. Cleanse the work as described in step no. 1, before painting.
- 3) Apply a thin coating of automotive primer paint. Auto-parts dealers sell this in handy spray cans. Allow the paint to dry for 12 hours, then apply a second coating. Set the work aside and allow the paint to dry for 24 hours. This may be hastened to two hours by placing the painted work in the oven. Set the temperature control to WARM.
- 4) The finish coat of paint may be applied next. Again, use spray-can paint, of your color choice. Avoid using bargain-price paint. The results may be disappointing with cheap paint. The nozzles often go bad, and blobs of dye tend to squirt out with the paint. This can negate all of your efforts to get the work ready for painting.
- 5) Work the spray can left and right as you paint. Keep the paint-can nozzle approximately one foot from your work to prevent excessive build-up and runs. Be patient! Apply a light layer of paint, wait 5 minutes, then apply a second thin layer, and so on. Normally, four or five thin layers of paint will suffice. Allow the work to dry in a dust-free environment for 24 hours.
- 6) A protective coating of polyurethane varnish may be added with a quality (fine bristle) brush if you wish a high gloss protective finish. This type of varnish is also available in spray cans if you wish a more perfect finish.

Adding Function Labels

Press-on decals are difficult to keep straight and level when you apply them. You need a steady hand and the patience of Job if you are to have results that look professional. I apply my decals before controls are mounted on the panel, and prior to attaching the panel to the chassis. This allows me to lay the panel on a flat surface and use a T-square for aligning the labels. It is vital that you wash all dirt and grease from your hands before starting this job. Grime will transfer from your fingers to transparent peel-off labels, and it will show after the label is affixed to the panel. This is not a major concern when you work with

press-on decals of the Datak brand variety.

After your labels are in place, lay a thin piece of paper over them and rub each label firmly with a smooth object, such as the rounded body of a plastic pen. This secures the labels to the panel. A coating of Krylon no. 1301 artist's spray (clear acrylic) or equivalent may now be added to help protect the labels. This protective spray is available at most artist and office-supply stores.

It is helpful to place the knobs, meters and other hardware on the panel before you commence adding your labels. This will prevent labels from being located too close to the panel holes. Otherwise, knobs and such may cover the labels when the equipment is assembled.

Plastic Dymo tape labels are effective for identifying the controls on your project. I keep a selection of tape colors on hand for this purpose. I use the same color paint as the label color, which tends to make the label less obtrusive when it is attached to the equipment. I prefer the smaller 1/4-inch wide tape to the rather overwhelming 3/8-inch tape.

An excellent label can be made with a Kroy lettering machine. Some artist-supply and office-supply stores have these machines. You can purchase custom-made labels for a modest fee. You can get white letters on a clear adhesive-backed tape, or you can obtain labels on clear tape with black letters. The chapter headings in this book were made with my Kroy model 80 machine. Various type sizes and styles are available through changing the Kroy print wheels as needed.

WIRE SELECTION and HINTS

Copper antenna wire is expensive and not always easy to locate if you live in a rural area. I'm referring to stranded no. 14 and no. 16 gauge conductor. Although this is an excellent antenna wire, you may use other types of copper conductor as well. Stranded wire is better than single-conductor wire because it can tolerate a lot of stress (wind-related movement) without breaking. Such wires as single strand no. 18 copperweld have good tensile strength, but break easily from flexing. This is a common problem when we use 450-ohm ladder line for feeding an antenna. This feed line contains no. 18 copperweld conductors, and they tend to break at the antenna feed point unless the polyethylene insulation below the attachment point is clamped to the feed point junction block.

I find that no. 18 stranded-copper speaker wire is entirely suitable for antenna and feeder use. The two conductors are encased in a clear vinyl material, and they pull apart easily. In effect, a 100-foot roll of this wire provides 200 feet of conductor, once separated. The plastic insulation adds strength to the antenna, but it is not unduly heavy. I find this kind of wire to be resistant to UV deterioration, and the insulation doesn't become brittle or crack when used out of doors. Speaker wire costs less than 5 cents per foot at current market prices.

If you wish to use no. 18 copperweld wire, don't overlook your farm-supply store as a source for this wire. It is sold in 1/4-mile spools for use in electric fence systems. I have paid as little as \$10 a roll for this wire.

Project Wire

Multiconductor telephone cable provides countless color-coded and insulated strands of single-conductor copper wire. This type of wire is suitable for short connecting

leads in small projects. I do not recommend this wire for mobile gear, since the vibration can stress these tiny wires to the breaking point. They should not be used for any lead that is subjected to frequent flexing. Scraps of this phone cable are often free for the asking if you know a phone installer that is willing to save a few pieces for you.

I prefer to use no. 22 or no. 24 stranded, insulated hookup wire for most of my projects. This wire is able to withstand a lot of stress and movement before the conductors break. I use this kind of wire also for making power-supply cables.

VHF Coil Wire

Stiff, self-supporting RF coils are often required when we build equipment for 28 MHz and higher. Tinned bus wire is hard to find. My inexpensive solution to this need is found by stripping the vinyl-plastic insulation from single strand no. 14 or no. 16 house wiring. I clean the copper with steel wool, then lay it in a tin-plating bath (see page 87). It is better to silver plate the wire in the interest of improved Q, but I experience fine results with tin-plated wire.

I use 1/8- and 1/4-inch copper tubing for some VHF applications. You can clean the tubing with steel wool and tin plate it to prevent discoloration and corrosion of the copper.

Double-sided glass-epoxy circuit board material may be used for constructing VHF and UHF strip lines. The copper on one side of the PC-board stock is joined to the copper on the opposite side at several points by soldering thin copper sheeting or brass shim stock to the conductors.

Ground Straps

You may remove the copper shield braid from RG-58, RG-59, RG-8 or RG-11 coaxial cable and use it as a ground strap from your station to the dc and RF grounding system. This conductor may also be tin plated to extend its life span in an outdoor environment.

Ground Radials

Care must be exercised when choosing the wire you plan to bury in the ground for an antenna ground screen. Some areas have soil that is high in acid or alkalinity. Buried wire, if bare, can disintegrate in a few months -- especially aluminum wire -- under these soil conditions. The heavier the wire gauge the longer it will last, but radials need not be of large cross section to work well. In theory, no. 30 wire is as satisfactory as no. 10 wire, since the radials carry very little RF current.

I prefer to use vinyl-covered no. 14 house wire for my buried radials. The ends of the wires are sealed with epoxy cement to prevent moisture and pollutants from migrating between the copper and the insulation. Enamel-insulated magnet wire is okay for soils of normal PH factor. I suggest that you use the type of enamel wire that is intended for a hostile environment of acids and oils. Formvar is one type of rugged enamel insulation. I have had the enamel coating vanish within 12 months when I buried it in highly acid soil.

Buried Coaxial Cable

The warnings offered in the previous section about soil acidity and alkalinity apply to buried coaxial cable as well. It is not uncommon to find that new RG-8, for example, becomes contaminated and lossy in a few months when it is buried in the ground. The ground pollutants cause the outer vinyl-plastic jacket to break down and release chemicals that spoil the insulating properties of the polyethylene inner insulation. When this occurs there is a power loss through the cable.

Impregnated 50-ohm cable is available. This material is electrically equivalent to RG-8, but has a sticky substance pumped into it. This compound resists damage from pollutants and prevents moisture from entering the cable. The outer jacket of the cable is UV-resistant, which makes it good for any outdoor application. The impregnant imparts a self-healing characteristic. This is helpful if rodents chew into the cable, which is not an uncommon event. This cable is sold as VB-8 (Decibel Products Corp., Dallas, TX) and as Impervon (Times Wire & Cable Co., Wallingford, CT).

Ordinary coaxial line that is used out of doors should always have weather seals at the open ends. This prevents moisture from entering the cable between the shield braid and inner insulation. Coax Seal, a putty type of commercial sealant, works well for this purpose. Quick-setting epoxy glue may also be used. Flexible caulking material (non-acetic) is a good material for sealing coaxial cable.

STANDARD COMPONENT VALUES

Capacitors, resistors and RF chokes are available in standard values. It is helpful to know these values when designing a circuit or placing a parts order. Table 4-3 lists standard RF-choke inductances.

Table 4-3

Microhenries (μ H)					Millihenries (mH)		
0.10	2.2	15.0	91	390	1.0	2.7	8.2
0.15	2.7	18.0	100	470	1.2	3.3	9.1
0.22	3.3	22.0	120	500	1.5	3.9	10.0
0.33	3.9	25.0	150	680	1.8	4.7	
0.47	4.7	27.0	180	750	1.8	5.6	
0.68	5.6	33.0	200	820	2.2	6.8	
0.75	6.8	39.0	220	910	2.5	7.5	
0.82	7.5	47.0	250	1000			
1.0	8.2	56.0	270				
1.2	9.1	68.0	300				
1.5	10.0	75.0	330				
1.8	12.0	82.0	350				

The choke values above are representative of those that are used by amateur experimenters. Millihenry values extend to 500 for commercial use. A complete listing of RF choke styles and values is provided in the J. W. Miller Co. catalog (19070 Reyes Ave., P.O. Box 5825, Compton, CA 90224).

Table 4-4

Silver Mica (pF)	Polystyrene (pF)	Disc (pF)
1	270	3.3
2	330	5
3	360	6
5	390	6.8
7	430	8
10	470	10
12	500	12
15	510	15
18	560	18
20	620	20
22	680	22
24	750	24
27	820	25
30	910	27
33	1000	30
36	1100	33
39	1200	39
43	1500	47
47	1600	50
50	1800	51
51	2000	56
56	2200	68
62	2400	75
68	2700	82
75	3000	100
82	3300	120
91	3900	130
100	4300	150
110	4700	180
120	5000	200
150	5100	220
160	5600	240
180	6200	10000
200	6800	
220	7500	
240	8200	
250		

Standard silver-mica, polystyrene and disc-ceramic capacitor values. Various dc-voltage ratings are available. For disc ceramic the range is 50 to 1000. The higher the operating voltage of all of the above capacitors the greater the physical size of the capacitor.

Standard values for computer-grade, electrolytic and tantalum capacitors are listed in **The ARRL Electronics Data Book**, second edition, 1988. Information concerning capacitor color coding is also available in that publication. The Data Book also contains a list of standard power-resistor values in ohms.

It would be a difficult task to compile an accurate and meaningful list of standard values for trimmer capacitors, since no two manufacturers seem to have the same capacitance ranges for their components. Generally speaking, these values are representative (specified in pF): 1-10, 1.5-15, 3-30, 4-60, 5-80, 10-100, 12-180, 25-280, 50-380, 80-480, 110-580, 140-680 and 170-780. Large mica trimmers are available also for pF ranges of 265-880, 340-1070, 425-1260 and 525-1415. Ceramic and plastic trimmer capacitors are available in the ranges below 10-100. Glass piston trimmers are made generally for a maximum capacitance of 10 pF.

Monolithic ceramic chip capacitors (for VHF and above) have values in the range listed for the disc-ceramic units in Table 4-4.

Resistors

Carbon composition and carbon film resistors are used most often in amateur work. 1/4- and 1/2-watt units are suitable for most solid-state circuits, although there are some applications that call for power ratings of 1, 2, 3 or 5 watts. These higher wattage ratings are necessary in a number of dc circuits. Noninductive power resistors (NIT) are used for ac and RF power circuits. Some hams use Globar resistors (noninductive) for RF applications, but the resistance value changes with heating (a design characteristic), and this makes them unsuitable for critical circuits.

Care must be taken when choosing the wattage rating of your resistors. If a given circuit causes a resistor to dissipate 1 watt, use a 2-watt or huskier resistor. You can calculate the power dissipation if you know the current that flows through the resistor, plus the voltage drop across it ($W = EI$). Under no circumstances should a resistor be more than warm to the touch after extended circuit operation. Power resistors should be mounted at least 1/4 inch above chassis and PC boards in order to provide reasonable air flow around the resistor body. They may be affixed to a heat sink by means of epoxy glue if greater cooling is needed. This is not recommended in circuits that have voltages greater than 100. Standard resistor values are listed in Table 4-5.

Table 4-5

1.0	7.5	56	430	3.3K	24K	180K	1.0M	5.6M
1.1	8.2	62	470	3.6K	27K	200K	1.1M	6.2M
1.2	9.1	68	510	3.9K	30K	220K	1.2M	6.8M
1.3	10.0	75	560	4.3K	33K	240K	1.3M	7.5M
1.5	11.0	82	620	4.7K	36K	270K	1.5M	8.2M
1.6	12	91	680	5.1K	39K	300K	1.6M	9.1M
1.8	13	100	750	5.6K	43K	330K	1.8M	10.0M
2.0	15	110	820	6.2K	47K	360K	2.0M	11.0M
2.2	16	120	910	6.8K	51K	390K	2.2M	12.0M
2.4	18	130	1000	7.5K	56K	430K	2.4M	13.0M
2.7	20	150	1100	8.2K	62K	470K	2.7M	15.0M
3.0	22	160	1200	9.1K	68K	510K	3.0M	16.0M
3.3	24	180	1300	10.0K	75K	560K	3.3M	18.0M
3.6	27	200	1500	11.0K	82K	620K	3.6M	20.0M
3.9	30	220	1600	12.0K	91K	680K	3.9M	22.0M
4.3	33	240	1800	13.0K	100K	750K	4.3M	
4.7	36	270	2000	15.0K	110K	820K	4.7M	
5.1	39	300	2200	16.0K	120K	910K	5.1M	
5.6	43	330	2400	18.0K	130K			
6.2	47	360	2700	20.0K	150K			
6.8	51	390	3000	22.0K	160K			

The resistors indicated in bold-face type are 10% values. Those in standard-weight type are 5% (gold band) values. All resistances are given in ohms. K = 1000 and M = 1,000,000.

NETWORKS AND ATTENUATORS

Attenuators, low-pass filters and broadband transformers are used frequently in amateur circuits. Basic information about these networks is worthy of inclusion in this chapter. The tabular data eliminates your need to do lengthy calculations when building a resistive attenuator or constructing a low-pass filter for use as a harmonic attenuator.

Resistive Attenuator Pads

Resistive attenuators are used to reduce signal levels while maintaining a 50-ohm circuit impedance. They are used also to ensure a 50-ohm termination for various circuits. Table 4-6 lists resistor values for pi and T networks that provide up to 40 dB of attenuation.

Table 4-6

PI ATTENUATOR			T ATTENUATOR		
<i>dB</i> atten.	<i>R</i> ₁ (ohms)	<i>R</i> ₂ (ohms)	<i>dB</i> atten.	<i>R</i> ₁ (ohms)	<i>R</i> ₂ (ohms)
1	870.0	5.8	1	2.9	433.3
2	436.0	11.6	2	5.7	215.2
3	292.0	17.6	3	8.5	132.0
4	221.0	23.8	4	11.3	104.8
5	178.6	30.4	5	14.0	82.2
6	150.5	37.3	6	16.6	66.9
7	130.7	44.8	7	19.0	55.8
8	116.0	52.8	8	21.5	47.3
9	105.0	61.6	9	23.8	40.6
10	96.2	70.7	10	26.0	35.0
11	89.2	81.6	11	28.0	30.6
12	83.5	93.2	12	30.0	26.8
13	78.8	106.0	13	31.7	23.5
14	74.9	120.3	14	33.3	20.8
15	71.6	136.1	15	35.0	18.4
16	68.8	153.8	16	36.3	16.2
17	66.4	173.4	17	37.6	14.4
18	64.4	195.4	18	38.8	12.8
19	62.6	220.0	19	40.0	11.4
20	61.0	247.5	20	41.0	10.0
21	59.7	278.2	21	41.8	9.0
22	58.6	312.7	22	42.6	7.8
23	57.6	348.0	23	43.4	7.1
24	56.7	394.6	24	44.0	6.3
25	56.0	443.1	25	44.7	5.6
30	53.2	789.7	30	47.0	3.2
35	51.8	1406.1	35	48.2	1.8
40	51.0	2500.0	40	49.0	1.0

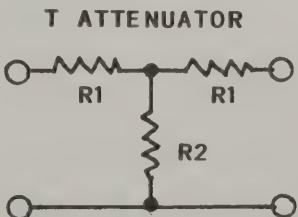
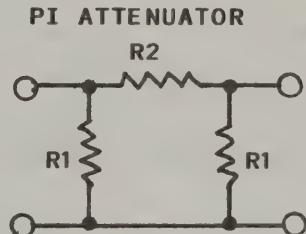


Table of resistance values for pi and T types of attenuators. The resistor wattage is selected in accordance with the power that is dissipated in the attenuator. Noninductive resistors are required. Equations for designing resistive attenuators for 50 ohms and other impedances are provided in The ARRL Electronics Data Book.

Broadband Transformers

The four most common broadband transformers for amateur work are shown in Figs 4-6 and 4-7. These are transmission-line transformers rather than conventional ones, the latter of which have primary and secondary windings that are separate from one another (not multifilar-wound). The transformers shown here are the bifilar and trifilar type. These transformers may be wound on toroid or rod cores. The permeability of the ferrite core is 850 to 900 for MF and HF applications. Amidon Assoc. no. 43 core mix is suitable. You may use 125 permeability cores for broadband transformers that operate from approximately 14 through 148 MHz (Amidon Assoc. no. 61 core material). Choose a core that can accommodate the anticipated power level. Cores may be stacked and glued together with epoxy cement to increase the effective cross-sectional area. An excellent text concerning the design and application of broadband transformers is **Transmission Line Transformers** by Jerry Sevick, W2FMI (an ARRL pub.). Also see **Ferromagnetic Core Design & Application Handbook** by Doug DeMaw (avail. from Amidon Assoc. or Prentice-Hall, Inc., Englewood Cliffs, NJ).

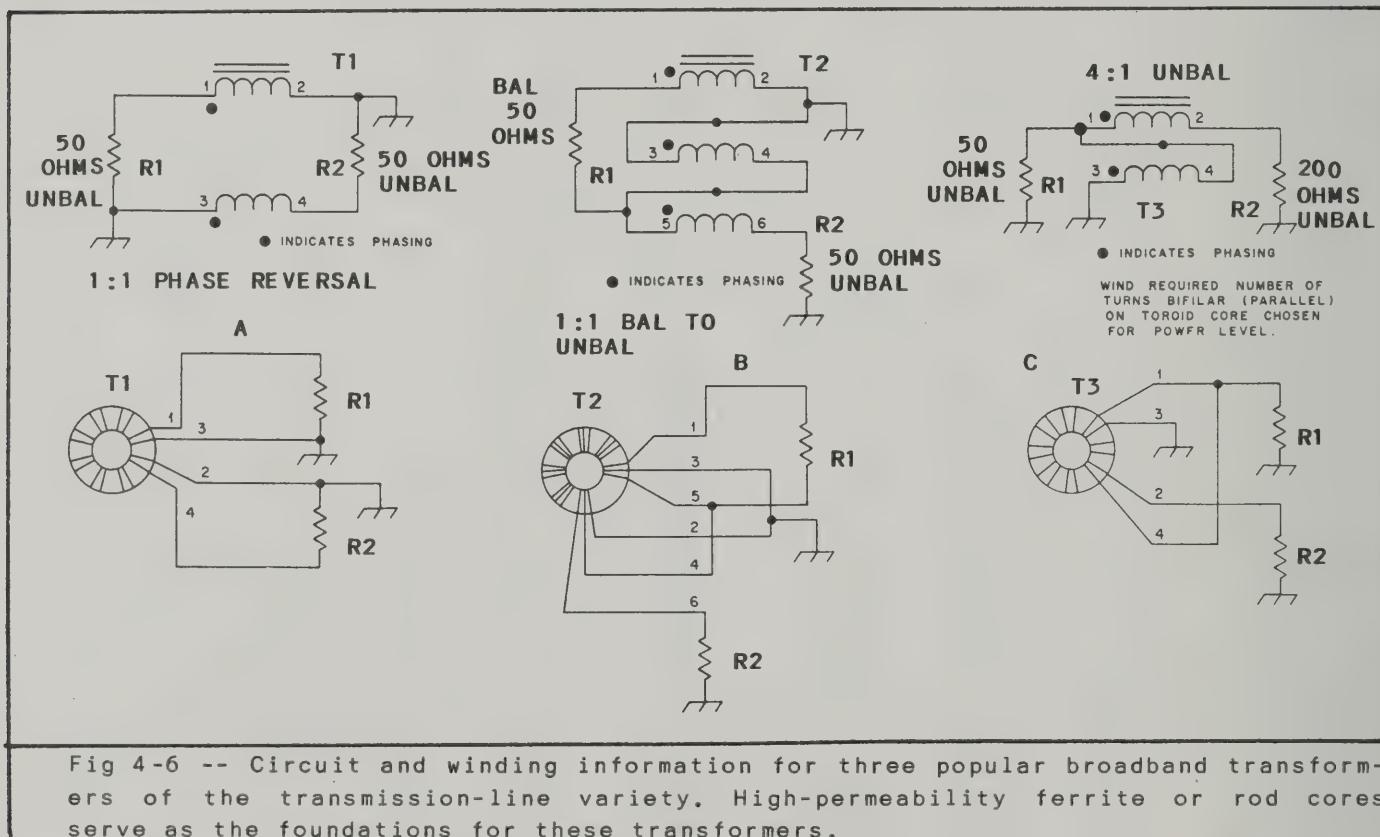
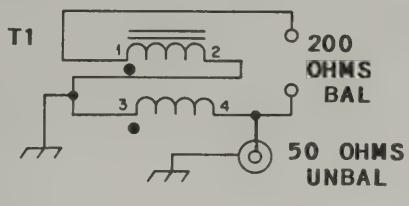


Fig 4-6 -- Circuit and winding information for three popular broadband transformers of the transmission-line variety. High-permeability ferrite or rod cores serve as the foundations for these transformers.

It is important to observe the phasing of the above transformer windings. If the phasing is not observed the transformer will not perform as specified. This job can be made easier if you use a different color of wire for each of the transformer conductors. The completed transformer may be dipped in polyurethane or glyptol varnish to protect it. This also keeps the windings from shifting on the core. I have used tool-handle dip as an encapsulant for broadband transformers that are intended for outdoor use. The clear dip is the least subject to deterioration from ultra-violet rays. This material is available in pint and gallon cans from United States Plastic Corp., 1390 Neubrecht Rd., Lima, OH 45801.



4:1 BAL TO UNBAL

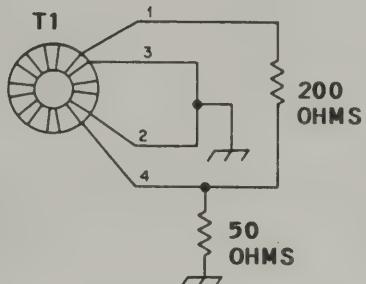


Fig 4-7 -- A 4:1 balanced to unbalanced broadband transformer. This configuration is used commonly for interfacing a Transmatch to balanced feeders, or to match a 50-ohm coax line to a balanced antenna that has a 200-ohm feed impedance. Suitable also for matching 75-ohm coax to a 300-ohm antenna, such as a folded dipole.

It is important to keep in mind that any broadband transformer should be used in an environment that has load impedances which match the transformation ratio of the transformer. These loads should be resistive rather than reactive in order for the transformer to function correctly. Broadband transformers are intended for use at impedance levels less than 500-600 ohms.

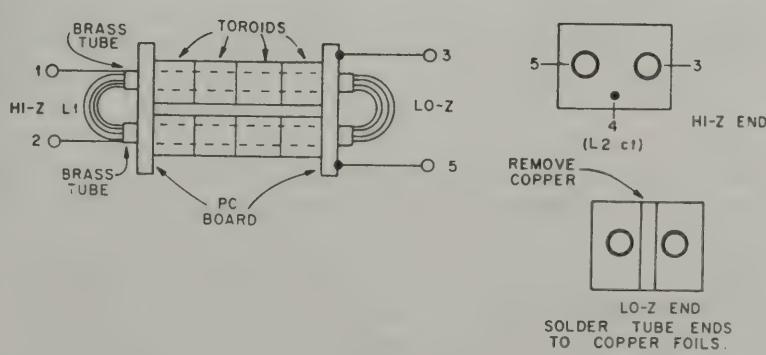
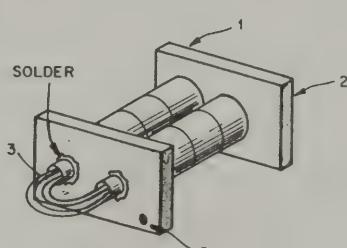
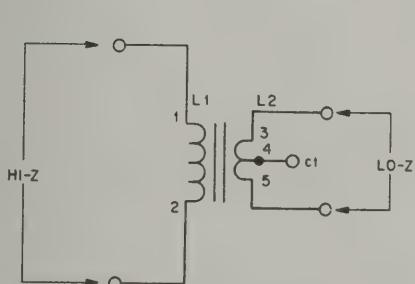


Fig 4-8 -- Illustration of a broadband transformer that may be used to match a solid-state power transistor to a 50-ohm harmonic filter. PC board material is used for the end plates. Two rows of ferrite toroids serve as the core material. Copper or brass tubing is used as a 1-turn primary winding. (Artwork courtesy of The ARRL.)

The transformer shown in Fig 4-8 is similar to the unit illustrated in Fig 4-1. They are electrically the same. However, the Fig 4-8 transformer is superior physically to the Fig 4-1E transformer. Mounting of the transformer is simpler, and the PC-board end plates provide a short path from the ends of the windings to the PC-board conductors. Stray inductance, however small it may be, can create matching problems at very low impedances. The stray inductance is reactive (XL). Capacitive reactance is sometimes introduced in a broadband transformer circuit in order to cancel unwanted inductive reactance.

Two lengths of copper or brass tubing (thin wall) are placed inside the rows of toroid cores. The end plate in Fig 4-8, numbered 3, 4 and 5, joins the two tubes. Their ends are soldered to the copper on the end plate. This forms an electrical and physical U. This conductor becomes a one-turn low-impedance winding for the transformer. The high permeability of the ferrite core ensures ample inductance for this winding. The high-impedance winding is looped back and forth through the tubing holes until the desired turns ratio is obtained. The end plate at the lower right in Fig 4-8 has the copped etched away at the middle. This creates two copper surfaces to which the tubing ends are soldered. These tabs may be soldered directly to the PC-board conductors. The two rows of cores may be cemented together, but this is not necessary.

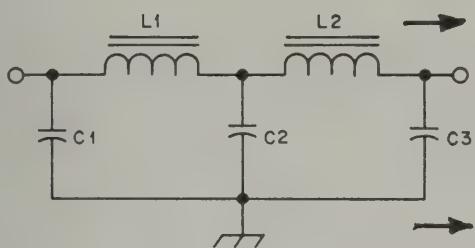
Harmonic Filters

Solid-state RF power amplifiers require a harmonic filter at the amplifier output. High-level harmonic currents occur in the collector circuit of RF transistors, and steps need to be taken to ensure acceptable spectral purity of the output signal. It is not uncommon to observe 2nd and 3rd harmonics that are only 10-15 dB below the fundamental energy at the output of an unfiltered amplifier. FCC regulations state that all spurious outputs must be 40 dB or greater below peak output power for MF- and HF-band operation.

The more filter sections or poles used the greater the harmonic attenuation. Low-pass filters are most often used for filtering the output of an RF power amplifier, although band-pass filters may also be used. There are many things to consider when selecting a filter, such as the SWR it creates in the line, plus its ripple factor. Ripple refers to humps and dips across the top or nose of the filter response. An ideal filter has a perfectly flat nose response, but this seldom occurs in practice. We can tolerate dips in the response if they are no deeper than, say, 1 dB below the flat part of nose.

The ARRL Handbook contains tables of normalized filter constants for designing high-pass and low-pass filters for 50-ohm loads. Various choices are available with regard to ripple factor and SWR. To use these tables all you need do is decide the filter cutoff frequency, then divide the frequency in MHz into the numbers listed in the tables for C and L. The values are in microhenries and picofarads.

Fig 4-9 shows two low-pass filters that you can use for filtering the output of a solid-state RF amplifier. The tables in Fig 4-9 are based on standard values of capacitance, as derived by Ed Wetherhold, W3NQN. The two-pole filter is acceptable for QRP transmitters that deliver less than 2 or 3 watts of output power. It is fine also for use in filtering low-level stages of a local oscillator or transmitter. The three-pole filter is Fig 4-9 is better in terms of harmonic attenuation. I recommend it, even for QRP applications.



BAND (m)	C1, C3 (pF)	C2 (pF)	L1, L2 (μ H)	f _{CO} (MHz)
80	560	1200	2.5	4.056
40	240	560	1.12	7.818
30	220	470	0.98	10.39
20	150	330	0.678	15.00
15	100	220	0.462	22.00

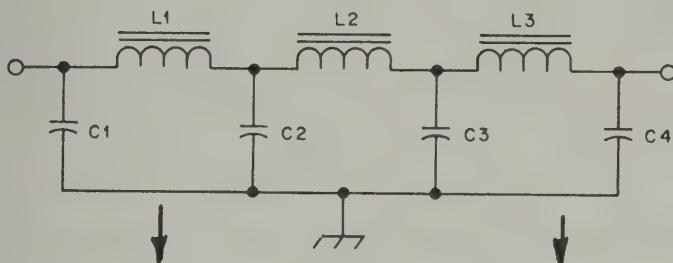


Fig 4-9 -- Data for two- and three-pole filters with a low-pass response. f_{CO} is the cutoff frequency where attenuation commences. Silver-mica or polystyrene capacitors are recommended.

BAND (m)	C1, C4 (pF)	C2, C3 (pF)	L1, L3 (μ H)	L2 (μ H)	f _{CO} (MHz)
80	510	1300	2.637	3.261	3.81
40	330	750	1.508	1.789	7.23
30	180	470	0.952	1.188	14.40
20	180	390	0.773	0.904	14.40
15	130	270	0.526	0.606	21.48
10	82	180	0.359	0.421	30.90

The filters in Fig 4-9 should use inductors that are wound on powdered-iron toroid cores. T37 (0.37 inch OD) cores are suitable for power levels up to 10 watts. Use T50 (0.5 inch) cores for power amounts up to 50 watts. T68 (0.68 inch) cores may be used for powers up to 100 watts. I recommend Amidon or Micrometals no. 2 core material (coded red) for frequencies up to 11 MHz. I prefer no. 6 (yellow) core material for 10 through 30 MHz. The proper choice of core mix ensures that the inductors have a loaded Q well above the filter design Q. Ferrite cores are not suitable for use in filters because they saturate easily and are unstable with regard to permeability versus heat.

The self-shielding characteristic of toroidal inductors helps to isolate the filter inductors. This reduces unwanted coupling between the input and output ports of the filter. Filter leakage (stray input-output coupling) can greatly reduce the overall harmonic attenuation of a filter. This is especially critical at the upper end of the HF spectrum, where small amounts of inductive and capacitive coupling become significant.

* INSERT APPROPRIATE DESIGNATIONS  METERS A-AMMETER V-VOLTMETER mA-MILLIAMMETER etc.		 MOTOR	 SINGLE CELL  MULTICELL BATTERIES	 CHASSIS  EARTH GROUNDS	 ASSEMBLY OR MODULE	 ANTENNA	 HEADSET	 QUARTZ CRYSTAL
COMMON CONNECTIONS  COAXIAL RECEPTACLE  FEMALE   	CONTACTS   	CONNECTORS    	 	  		   		
 	  	  	  	  	 			
  	  	  	  	 	 	 	 	
 	 		 	 	 	 	 	 
 	 	 	 	 	 	 	 	 

Fig 4-10 -- Standard EIA electrical symbols that are used in this book. Publishers of other technical books may not use these symbols for reasons of style preference.

PARTS PROCUREMENT

I have heard countless amateurs say, "I'd build equipment if only I could find the parts for the projects." Frankly, that is not a good reason to avoid the ham workshop. Parts are more abundant and inexpensive than ever before -- at least with respect to solid-state circuits. The lament may be true, indeed, when we consider availability and cost for vacuum tubes, large inductors and variable capacitors. As the demand for these parts decline, up goes the price. But, for the solid-state enthusiast, things couldn't be better.

Sources for Parts

The best arena in which to stalk those fugitive components is the ham radio flea market. A few dollars and a gunnysack will equip you for a productive sortie among the hawker's display tables. We hams tend to buy only those parts for a project of immediate interest. If you're an experimenter it is better to stock up on common parts that you will need for future projects. Such items as switches, meters, resistors, capacitors, semiconductors and blank PC boards are among the things I like to horde. It's joyful to reach in the parts drawers and pull out the components I need for experimenting. Don't be timid when you see a good deal at a swap-and-shop session!

Surplus PC boards and equipment are often available at rock-bottom prices. These items contain a wealth of reuseable parts if you're willing to strip the goodies and store them for use later on. Cast-off transistor radios and TV receivers also yield many excellent parts for ham projects.

Surplus Parts Vendors

It seems that some new surplus dealer pops up each day. Existing ones tend to wax and wane, which makes it difficult for me to offer an active list of names and addresses for these dealers. The following run-down represents dealers from whom I buy many of my components at the time this book is being written. Each dealer in the list offers a catalog. It is helpful to enclose \$1 with your catalog request. This helps to offset the dealer's cost for having the catalog printed.

All Electronics Corp. P.O. Box 567 Van Nuys, CA 91408 1-800-826-5432	Mouser Electronics 2401 Hwy 287 N. Mansfield, TX 76063 1-800-346-6873	Circuit Specialists P.O. Box 3047 Scottsdale, AZ 85271 1-800-528-1417
BCD Electro P.O. Box 450207 Garland, TX 75045-0207	Fair Radio Sales Co. P.O. Box 1105 Lima, OH 45802	Mid-America Corp. 2309 S. Archer Ave. Chicago, IL 60616 1-800-621-1530
Digi-Key Corp. 701 Brooks Ave., S. P.O. Box 677 Thief River Falls, MN 56701-0677 1-800-344-4539	Jameco Electronics 1355 Shoreway Rd. Belmont, CA 94002	Oak Hills Research 20879 Madison Ave. Big Rapids, MI 49307 (616)796-0920 Send large s.a.s.e. w/50¢ postage for catalog.
Marlin P. Jones & Assoc. P.O. Box 12685 Lake Park, FL 33403-0685	Hosfelt Electronics, Inc. 2700 Sunset Blvd. Steubenville, OH 43952 1-800-524-6464	

I wish to stress the "caveat emptor" (let the buyer beware) principle. Although I have always found the dealers in the foregoing list to be entirely reputable, the ARRL, Inc. and I are in no way responsible for the quality of the service and merchandise of these dealers.

Suppliers of Other Materials

We amateurs need a number of products that aren't available from most surplus vendors. Hardware, etched and drilled PC boards for **QST** and **Handbook** projects, plus plastic materials. The following list of dealers is worth adding to your roster of names. Be sure to request a catalog.

ETCHED & DRILLED PC BOARDS

A & A Engineering
2521 W. La Palma Ave.
Unit K
Anaheim, CA 92801

Circuit Board Specialists
P.O. Box 951
Pueblo, CO 81002

FAR Circuits
18N640 Field Court
Dundee, IL 60118

TUBING & HARDWARE

Small Parts, Inc.
6891 N.E. Third Ave.
P.O. Box 381966
Miami, FL 33238-1966

PLASTIC TUBING, RODS & SHEETS

United States Plastic Corp.
1390 Neubrech Rd.
Lima, OH 45801
1-800-537-9724

TOROIDS, RODS & POT CORES

Amidon Associates, Inc.
12033 Otsego St.
N. Hollywood, CA 91607

Toroids are available also from Palomar Engineers, Box 455, Escondido, CA 92025. A large list of RF power transistors is available from RF Parts, 1320 Grand Ave., San Marcos, CA 92069. Try to keep a close watch on the ads in **QST**. Many parts and equipment suppliers use the pages of **QST** for their advertising.

Chapter Summary

I have tried to provide most of the information that experimenters need to plan and construct projects. The success of your workshop efforts depends in large measure upon your initiative toward collecting catalogs and parts. The various diagrams and tables in this chapter have been included so that you can avoid investing your time in calculations and related reference work. Be sure to check the expanded listing of parts suppliers in the Component Data chapter of the latest editions of **The ARRL Handbook**.

PRACTICAL RECEIVERS & TECHNIQUES

CHAPTER 5

Perhaps the most fascinating part of our hobby is associated with the design and building of receivers. Many of us experienced the thrill of hearing a crystal-clear AM broadcast signal when we built our first crystal detector radio with a galena crystal and a cat's whisker electrode. This crude form of receiver was commonly available in kit form at "five and dime" stores when commercial broadcasting commenced many decades ago. All that was needed was a tuned circuit, a pair of earphones, a long piece of wire for an antenna, and a good earth ground. The higher the Q of the tuned circuit the greater the receiver selectivity for separating the stations. Unfortunately, a strong, local AM station tended to overwhelm these receivers, and this prevented us from copying the weaker stations that were nearby in frequency. The galena crystal (carborundum was used also) and its cat-whisker contactor formed a semiconductor diode junction, which in turn converted the RF (ac) signal to pulsating dc, and this was audible in the earphones in the form of voice and music energy. The users had to keep adjusting the position of the cat's whisker in order to find a "hot spot," as they were called. The hot spot provided the loudest signal in the earphones.

We have come a long way since the days of the crystal detector. Vacuum tubes were developed and the TRF (tuned radio frequency) radio came into being. Several tuned circuits were employed in order to achieve sufficient selectivity to separate the many AM stations from 540 to 1600 kHz. One or more RF amplifier stages were used ahead of a vacuum-tube detector. The audio signal was then amplified with more tubes to provide headphone or loud-speaker volume.

Finally, the superheterodyne radio was conceived. It provided excellent gain and selectivity, and it has remained the standard circuit to date. Modern radios have reached a level of performance and sophistication that was only a notion in the early 1930s. Analog frequency readout is fast becoming a method of the past as the newer receivers become equipped with digital frequency displays. The ability of a receiver to withstand the onslaught of strong signals (dynamic range) without performance degradation is a design criterion for most of the dedicated manufacturers. Frequency stability is another design objective. This has been satisfied in a satisfactory manner through the introduction of PLLs (phase-locked loops) and synthesizers. Fewer and fewer LC (inductance-capacitance) local oscillators or VFOs are being used in receivers and transmitters, owing to their tendency to drift.

The Simplest Receiver

You will enjoy constructing a crystal-detector radio if you have never tried your hand at this facet of experimenting. A practical circuit that tuned from 540 to 1600 kHz is presented in Fig 5-1. No cat's whisker is needed because we are using a modern germanium diode as the detector. The circuit can be assembled

on a piece of wood, or you may want to build it on one of the universal breadboards described on page 88. The most tedious part of this project is the winding of L2. You may wind the coil in increments. Take a break periodically and anchor the already-wound turns temporarily with a piece of adhesive tape.

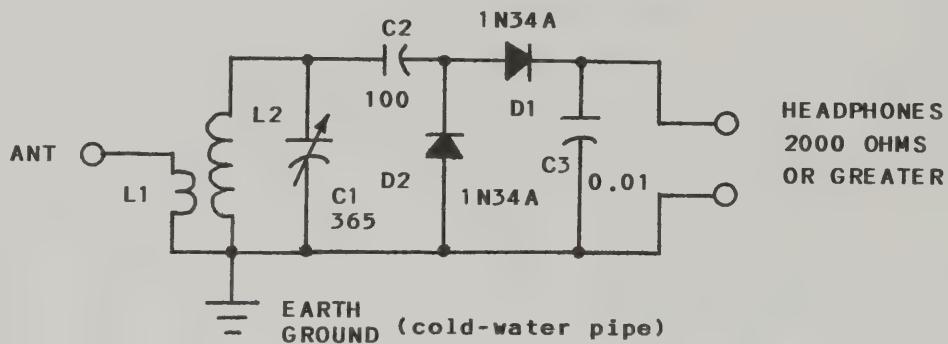


Fig 5-1 -- Practical circuit for a simple broadcast band crystal-detector receiver. D1 and D2 are detectors that operate as a voltage doubler to provide greater headphone volume. C1 tunes L2 to the desired frequency. C1 is a 365-pF variable capacitor (Mouser Electronics no. 24TR218 suitable. Use both sections in parallel.). L2 is wound on the cardboard tube from a roll of toilet tissue (1-1/2 inch dia) and occupies 3-3/4 inches of the tube. Use 168 turns of no. 26 enamel wire, close wound, to obtain 380 uH of inductance. Other coil forms may be used. L1 contains 20 turns of no. 26 enamel wire, close wound over the grounded end of L2.

Modern-day low-impedance headphones are not suitable for use with the circuit above. You will need a pair of the older high-impedance phones in order to obtain ample volume. You may wish to add a single transistor audio amplifier at the receiver headphone terminals. This will permit the use of 8-ohm phones. See Fig 2-1A for an example of a suitable audio amplifier. C1 and C2 in Fig 5-1 are in pF. C3 is in uF.

Regenerative Receivers

One- and two-tube "gennys" or regenerative receivers were popular in the early days of Amateur Radio. The circuit was constructed to permit self-oscillation of the detector stage by means of a regeneration (feedback) control. This control was adjusted so that the detector was oscillating weakly (for CW reception) and it was set just at the brink of oscillation for AM reception. This was the most sensitive condition for the detector. A regenerative receiver may be used to copy SSB and CW signals by allowing the oscillating detector to create the needed beat note for CW and the carrier for the SSB energy. This simple type of receiver should not be overlooked if you wish to build a small portable receiver with minimum components.

A bad feature of "regens" and DC (direct conversion) receivers is that they radiate a weak signal on the frequency to which they are tuned. If you live near another ham and monitor his QSO, chances are that the energy from your receiver will interfere with his reception. This unwanted radiation can be minimized if an RF amplifier is used ahead of the detector.

There are a number of ways to control self-oscillation in a regenerative receiver. The simplest method is to vary the operating voltage of the detector. Better performance generally results if the regeneration control is arranged to change the amount of feedback energy in the detector. Fig 5-2 shows the circuit of a solid-state regenerative receiver.

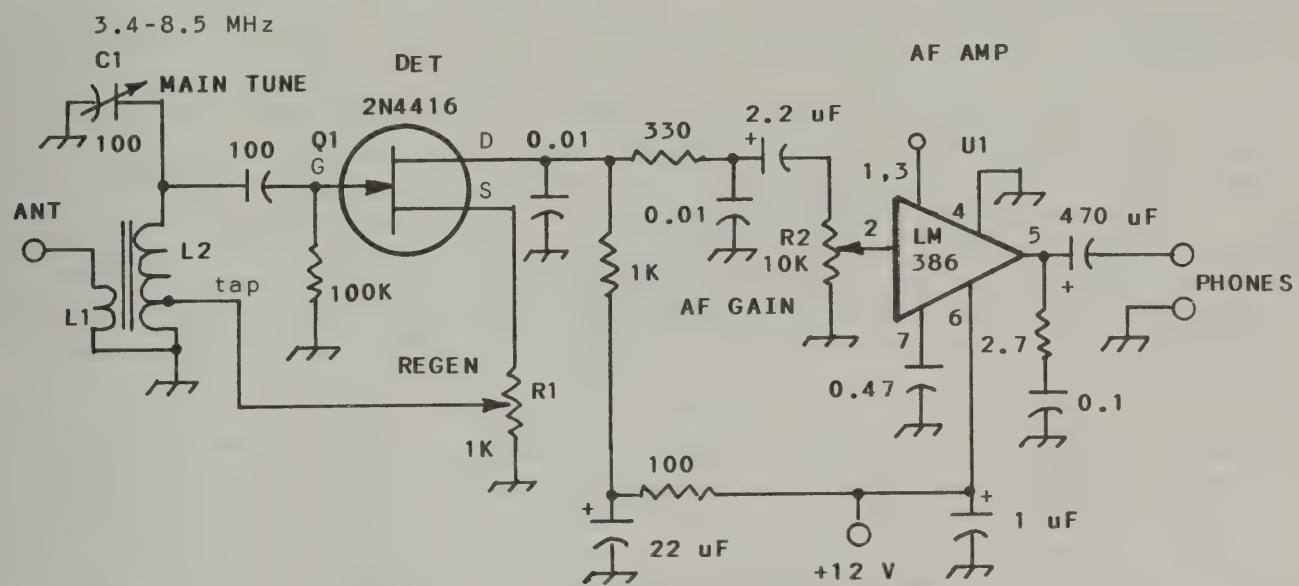


Fig 5-2 -- Practical circuit for a two-stage regenerative receiver. Decimal-value capacitors are in μ F. Others are in pF. Polarized capacitors are tantalum or electrolytic, 16 V or greater. Fixed-value resistors are 1/4- or 1/2-W carbon film. R1 is a linear taper carbon-composition control. R2 is an audio taper carbon-composition control. L2 has 22 μ H of inductance to cover 3.4-8.5 MHz with C1. Use 46 turns of no. 26 enam. wire, close wound, on a 1-inch dia X 2-inch coil form. Place tap at 11 turns above the grounded end of the coil. L1 has 5 turns of no. 26 enam. wire, close wound, over the grounded end of L2.

Regenerative receivers are subject to the effects of hand capacitance unless care is taken when planning the receiver layout. Incorrect layout and assembly will lead to shifting of the frequency when your hand is placed near the panel of the receiver. In the days of yore we often joked about the need to tune our gennys with a wooden broom handle in order to avoid hand-capacitance effects. It is wise to use a metal panel on the receiver. R1 needs an insulated tuning shaft brought to the front panel. This is because it is "hot" with feedback energy. A piece of 1/4-inch wooden dowel rod may be used as an insulated shaft.

The receiver may be fine tuned by placing a 10-pF variable capacitor in parallel with C1. In doing this you cause C1 to function as a bandset control, while the 10-pF variable capacitor acts as a bandspread control. A vernier drive can be used for the bandspread capacitor to make tuning easier. Other frequency ranges are possible by simply changing the inductance of L2. The same tap and turns ratio for L1 and L2 should be maintained if this is done.

If you construct a set of plug-in coils for the Fig 5-2 circuit, you can listen to many MF and HF bands by simply changing coils.

You are probably wondering why I bothered to discuss crystal detectors and regenerative receivers. After all, they are somewhat archaic by today's standards. First, I want you to be familiar with basic receiver circuits and secondly, I wish to encourage you to experiment with both circuits as part of a learning exercise. The business of learning "how to crawl before learning to walk" has merit for the neophyte experimenter, and that's what this book is all about. The Fig 5-2 receiver can be built easily on the universal breadboard.

DC Receivers

DC in this example means "direct conversion" rather than direct current. There has been some confusion about this term. Some amateurs refer to DC receivers as homodyne or synchrodyne receivers. In effect, the receiver detector (front end) is a product detector that is tunable. It requires the injection of a VFB0 (variable frequency beat oscillator) in order for it to function as a detector for CW and SSB. The output from the detector is at audio frequency. This type of receiver has become a favorite among hams who build portable QRP equipment. It appeals also to those who wish to construct simple but effective ham-band receivers.

The shortcomings of the DC receiver have been outlined in **The W1FB QRP Notebook**, but they are worth repeating here. (1) Susceptibility to unwanted AM detection. (2) Subject to common-mode hum caused by VFB0 energy entering the ac power supply diodes and being reradiated with 120-Hz hum modulation. (3) They require 80-100 dB of audio amplification to exhibit good overall sensitivity. The high audio amplification causes front-end mechanical noises to be amplified. These noises appear as microphonics when the receiver is bumped.

The good circuit features should be discussed also. (1) Minimum component count. (2) Easy and inexpensive to construct. (3) They do not require hard-to-find parts. (4) They do not suffer from "dead spots" in the tuning range, which is the case with a regenerative receiver.

How might we defeat some of the problems that are attendant to DC receivers? First, microphonics can be reduced if we use an RF amplifier or preselector ahead of the detector. This additional gain (10 to 25 dB, depending on the preamp design) can be subtracted from the gain of the audio-amplifier channel. However, the greater the gain ahead of the detector the worse the receiver DR (dynamic range). This is true also of superheterodyne receivers.

Common-mode hum can be minimized by using an RF choke in each of the power supply output leads, directly at the power-supply terminals. Each diode rectifier in the power supply should be bypassed with a 0.01-uF disc ceramic capacitor. A good earth ground on the DC receiver helps to lessen the effects of common-mode hum. This is not a problem when operating a DC receiver from a battery.

Unwanted AM detection (interference and blanketing from commercial AM shortwave stations) can virtually be cured by using a doubly balanced product detector (four diodes or an active detector with an IC).

Another DC Receiver Shortcoming

Single-signal reception is not possible with a DC receiver. The system responds to signal energy that is either side of zero beat. For example, if you are listening to a CW signal that corresponds to USB (upper sideband), and if there is

another signal located on the opposite side of zero beat (equivalent to LSB in this case), it will seem that the unwanted signal is on or near the frequency of the desired one. The interfering signals would not be heard if you were using a superheterodyne receiver of good design. This performance tradeoff is acceptable to most experimenters in return for circuit simplicity and otherwise good performance.

The addition of a passive LC or RC active band-pass filter in the audio channel will lessen the effects of QRM significantly with regard to signals that are above and below the operating frequency.

DC Receivers

There are countless designs for DC receivers. You will want to obtain copies of **The W1FB QRP Notebook** and **Solid State Design for the Radio Amateur**. Both books contain many DC receiver circuits, along with design data. A simple but practical circuit example is provided in Fig 5-3.

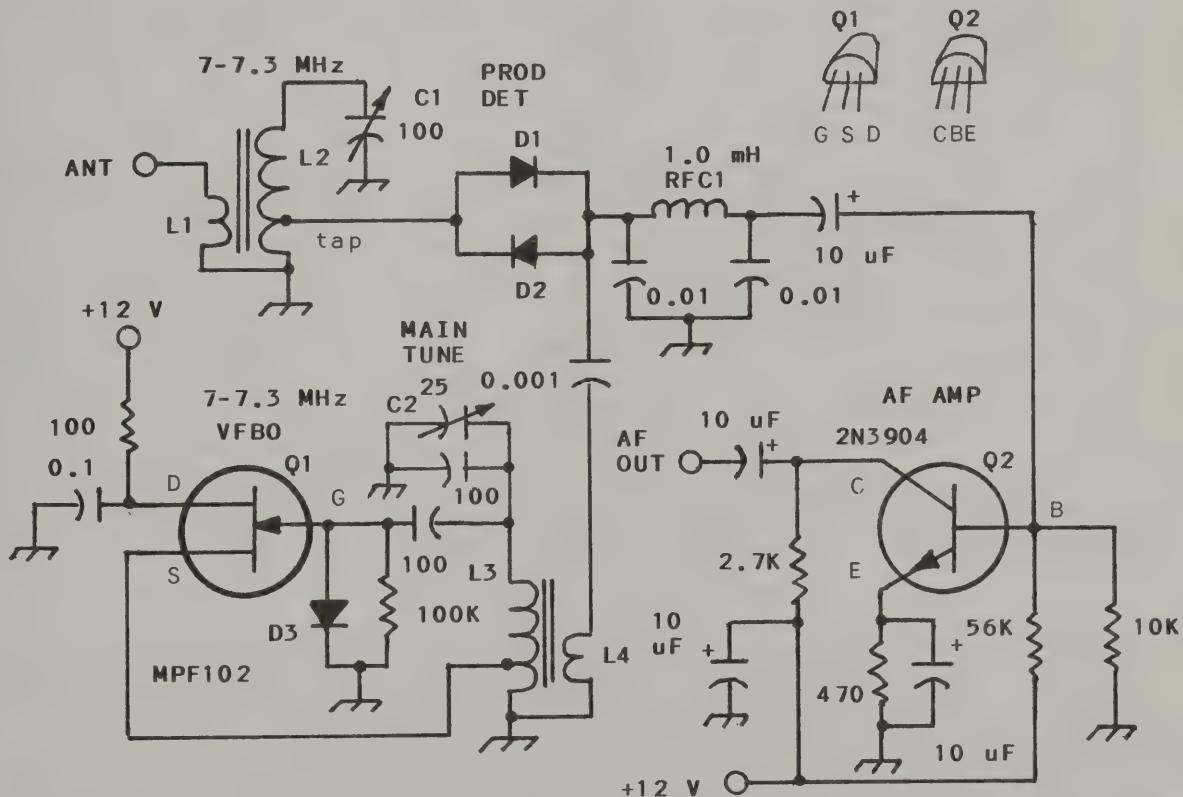


Fig 5-3 -- A simple DC receiver for 40 meters. C1 is a 100-pF trimmer. C2 is a 25-pF miniature air variable with vernier drive. D1 and D2 are matched (preferably) 1N914 diodes. D3 is also a 1N914. L2 is a 9-uH toroidal coil (40 turns of no. 26 enam. wire on an Amidon T68-2 toroid). L1 has 4 turns of no. 26 enam. wire. L3 is a 4.8-uH coil (32 turns of no. 26 enam. wire on an Amidon T68-6 toroid). L4 has 6 turns of no. 26 enam. wire. Polarized capacitors are tantalum or electrolytic. Decimal-value capacitors are in uF. Others are in pF. Resistors are 1/4-W carbon film. Tap L2 at 5 to and tap L3 at 8 turns.

The local oscillator (VFBO) in a DC receiver operates at the receive frequency, as indicated in Fig 5-3. The VFBO is offset from the incoming signal frequency sufficiently to produce the desired beat note for CW (500 to 1000 Hz typically). The VFBO is adjusted during SSB reception to provide the missing carrier for the USB or LSB signal. The product of the incoming signal and that of the VFBO is, therefore, an audio frequency. A regenerative receiver functions in a similar manner, except that the detector self-oscillates to provide the equivalent of the VFBO output frequency.

The detector (D1 and D2) in Fig 5-3 was developed by V. Polyakov, RA3AAE, and was described in **Radio** for December 1976. It is intended to provide an alternative to a doubly balanced diode detector. However, the configuration shown in Fig 5-3 does not match the performance of a diode-quad doubly balanced mixer. An improved version of this circuit that uses four diodes was described in the RSGB book entitled **Amateur Radio Techniques**, 7th edition, page 131.

RFC1 in combination with the two 0.01-uF capacitors acts as an RF filter to keep the VFBO energy out of the AF amplifier, Q2. Smaller values of capacitance in the filter will provide greater high-frequency response in the audio channel. One stage of audio amplification is shown. This is adequate for experimenting with this circuit, but two more identical stages of amplification are necessary in order to ensure adequate weak-signal reception.

The VFBO circuit may be improved by regulating the Q1 drain voltage with a 9.1-volt, 400-mW Zener diode. The Fig 5-3 circuit is included primarily to illustrate in a simple manner how a DC receiver operates. I recommend that you construct the circuit for the purpose of becoming familiar with how DC receivers perform. The circuit is entirely adequate (with added audio amplification) for use in simple emergency and portable transceivers. Best VFBO stability will result if the two 100-pF capacitors are NPO ceramic types. L3 should be treated with three coatings of General Cement Polystyrene Q Dope or equivalent high-Q cement. This will keep the coil turns from shifting and causing frequency changes. A 2N4416 JFET will provide better overall VFBO performance than will an MPF102. It has a higher transconductance (better for oscillation) and a better pinchoff characteristic (allowing greater power output). Hot-carrier or Schottky diodes at D1 and D2 will also lead to improved performance.

An Improved DC Receiver

The Signetics NE602 mixer IC is a low-cost chip that offers good performance as a product detector or mixer. It was popularized in **QST** for February 1988 ("The Neophyte Receiver") by D. Dillon, WA3RNC. I recommend that you read his article and duplicate the receiver. It is a monument to simplicity, but it offers good performance for the small parts count he specified.

Although the NE602 is designed with an internal oscillator, requiring only a few external LO components, I prefer to use a separate LO or VFBO. This is because all oscillators contain a substantial amount of harmonic current, and this energy leads to unwanted injection frequencies that cause spurious responses. When the active devices for an oscillator are on the same substrate as the other IC elements, it is impossible to isolate the harmonic currents. If we use an outboard LO we do not encounter this problem. However, for simple receivers intended for casual use, the NE602 may be used as an oscillator and mixer or PD without serious ills.

Fig 5-4 shows the circuit for a DC receiver that uses the NE602 IC. It lacks some of the refinements that are found in high-performance DC receivers (such as a three-pole RC active filter), but it gives a good account of itself for general amateur use.

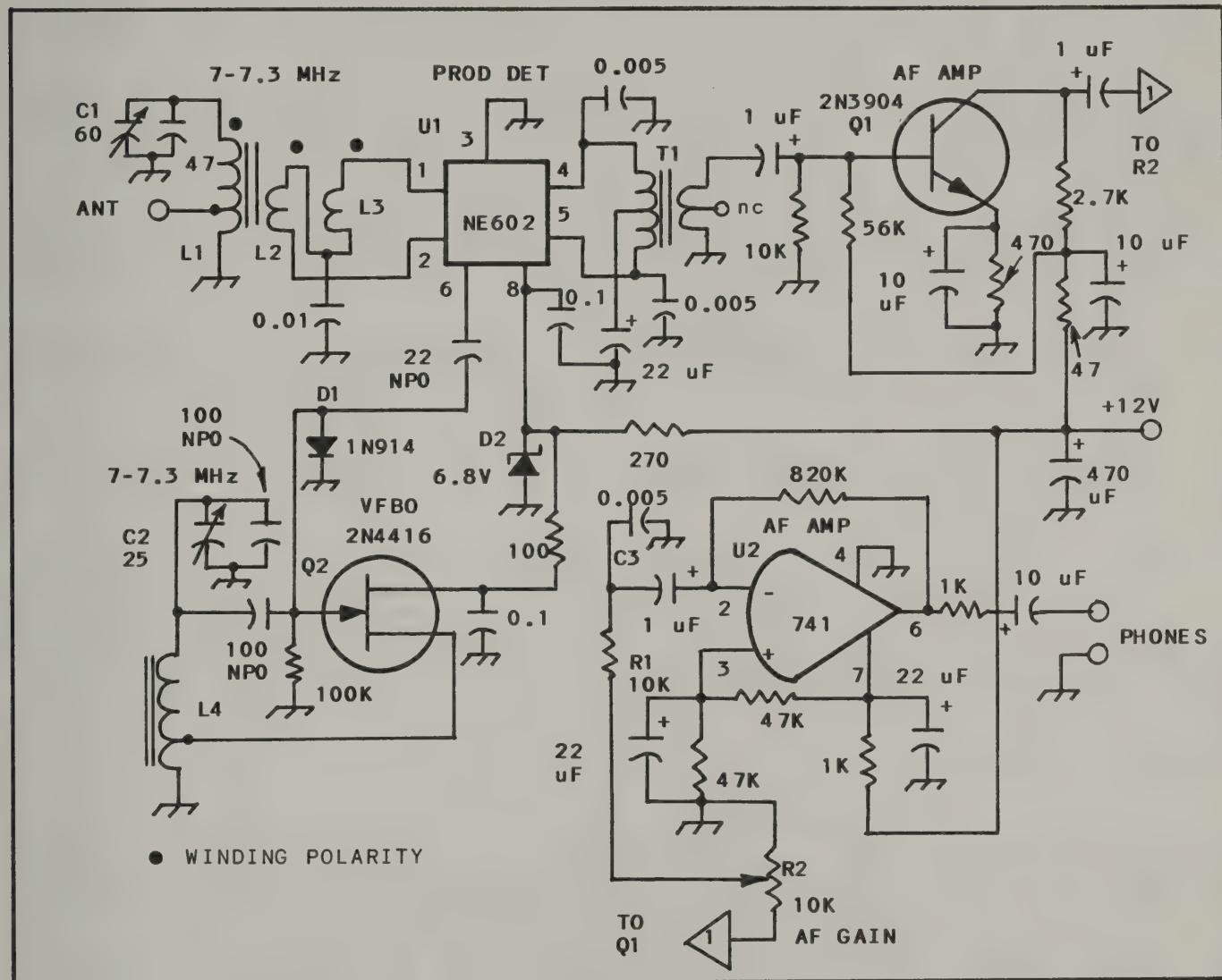


Fig 5-4 -- Schematic diagram of a 40-meter DC receiver. Operation on other bands can be achieved by approximately doubling the tuned-circuit values for 80 m or halving them for 20 m. Fixed-value capacitors are disc ceramic. Polarized capacitors are tantalum or electrolytic, 16 V or greater. Fixed-value resistors are 1/4-W carbon film. C1 is a 60-pF trimmer and main-tuning capacitor C2 is a 25-pF miniature air variable (plus vernier drive). L2 is 9 μ H. Use 40 turns of no. 26 enam. wire on a T68-2 toroid. Tap 4 turns above grounded end. L2 and L3 are 10 bifilar turns of no. 26 wire over L1 winding (observe polarity). L4 is 4.8 μ H (32 turns of no. 26 enam. wire on a T68-6 toroid. Tap at 8 turns above grounded end. T1 is a 4K-ohm ct to 600 ohms ct audio transformer (Mouser 42TL021 or equiv.). R2 is an audio-taper, 10K-ohm carbon-composition control. Polystyrene capacitors may be substituted for the NPO capacitors.

The Fig 5-4 receiver lacks a high degree of selectivity. R1 and C3 have been included to form a simple low-pass filter that has approximately 16 dB of attenuation at frequencies above roughly 800 Hz. The primary of T1 is tuned by means of the two 0.005- μ F bypass capacitors. It has a band-pass response in the audio range of interest. If you use a different interstage transformer, such as a 10K-ohm to 2K-ohm type, measure the primary inductance and use two capacitors that provide a peak audio response at 1000 Hz. This improves the overall selectivity of the receiver.

The VFBO injection to pin 6 of U1 should be 200 mV P-P or greater, but not greater than 500 mV. Measure the pin 6 voltage with a scope and select a capacitor value (22 pF in Fig 5-4) that provides roughly 350 mV P-P at pin 6.

C1 is adjusted for peak signal response at 7.1 MHz. You may wish to add a 25-pF NPO trimmer capacitor in parallel with C2 to facilitate calibrating the main-tuning dial of the receiver.

An RC active audio filter may be included to enhance the receiver selectivity. It should be inserted between Q1 and U2 if this is done. Eliminate R1 and C3 if you use an active filter.

An LM386 audio IC can be added after U2 if you wish to use a loudspeaker. See Fig 5-2 for the circuit which uses an LM386 chip.

The Fig 5-4 caption indicates that this receiver may be used on other HF bands by changing the tuned-circuit constants. I do not recommend that you use this circuit above 14 MHz. This is because oscillator stability becomes difficult to achieve above 20 meters. Also, since the noise figure of the NE602 is fairly high (12 dB), a low-noise preamplifier should be used ahead of U1 for 20-meter operation. It is not required for frequencies below 14 MHz.

AGC for the DC Receiver

Automatic gain control (AGC) is a desirable feature in any receiver. Without it we are subjected to ear discomfort when we tune across a very loud signal, or when one of these whopper signals suddenly appears on or near the frequency of interest. This annoyance is compounded by the use of earphones. Various methods for applying audio-derived AGC have been developed. Modern IC technology makes it a simple matter to use an audio compressor in a DC or superheterodyne receiver at minimal cost. A compressor functions somewhat like an AGC circuit.

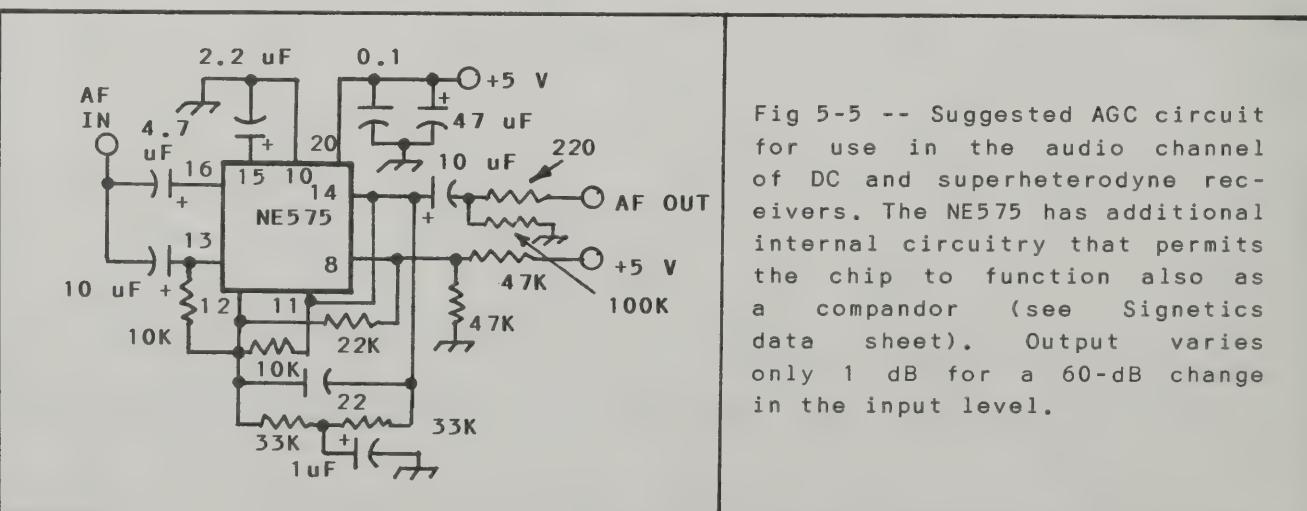


Fig 5-5 shows a Signetics circuit that is designed to operate as an ALC (automatic limiting control) system. Irrespective of the audio-input level over a 0 to 60-dB range, the output changes by only plus or minus 0.5 dB. This type of circuit lends itself well to AGC use in simple receivers that are not capable of providing an IF signal for IF-derived AGC. The NE575 can be used as an ALC or compandor chip, depending upon how it is configured. It can be made switchable to provide both functions.

Full audio-derived AGC can be had by sampling the AF energy ahead of the AF gain control, amplifying the sampled audio, then rectifying and filtering it. The developed dc voltage is then used to reverse bias one or more of the low-level AF stages ahead of the AF gain control. Normally, a 0.5 to 1 second decay time is provided for the AGC action. A simple parallel RC network may be shunted to ground from the rectified AF line to provide this time constant. A S meter may be operated from the AGC line if you desire a relative indication of the strength of incoming signals. Fig 5-6 shows an audio-derived AGC circuit.

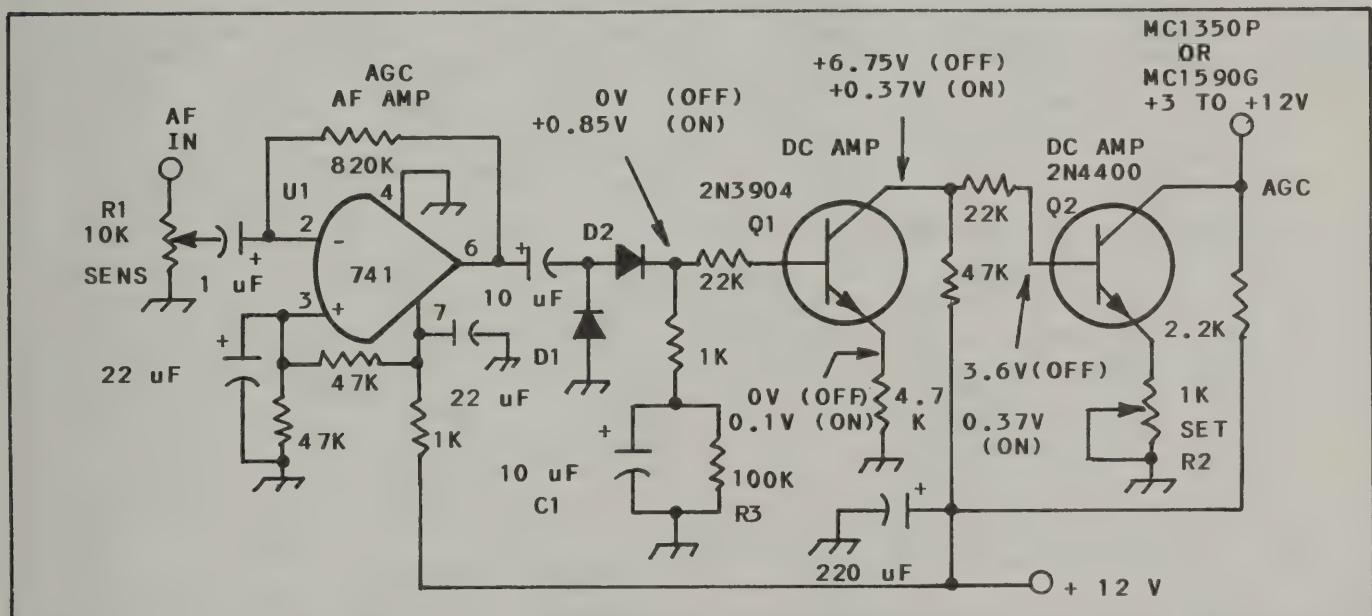


Fig 5-6 -- A practical audio-derived AGC circuit for use in a superhet receiver that uses MC1590G and MC1350P ICs as IF amplifiers. C1 and R3 establish a 1 second AGC delay time. A tantalum capacitor is best for C1. D1 and D2 are 1N914 or equivalent silicon diodes. Maximum IF gain occurs when the AGC voltage is +3. The greater the AGC voltage the lower the MC-1590G or MC1350P gain. R1 is adjusted for only enough audio-input voltage to produce the dc voltages marked on the diagram. R2 sets the AGC output-voltage range at +3 to +12 volts. OFF is when no audio input voltage is applied to pin 2 of U1. ON indicates full AGC action.

I have used the Fig 5-6 circuits a number of times. The major limitation is the somewhat nonlinear voltage change at the collector of Q2. In other words, the level change is gradual at low AF input levels, but is rapid over the upper 2/3 of the AGC range. A PNP transistor may be substituted at Q2 to obtain an AGC that is low at high input-signal levels and high at low input-signal levels. The Q2 resistance values may require modification to obtain the desired AGC range. A PNP transistor at Q2 makes this circuit suitable for using the AGC voltage to control a low-level audio-amplifier stage rather than an IF amplifier.

Superheterodyne Receivers

It is interesting that we have moved forward from the crystal detector to the TRF radio, then on to the regenerative receiver and finally, the superheterodyne circuit, without any further change in the general concept of the superhet. Many circuit refinements, such as high orders of selectivity, excellent AGC, double and triple conversion schemes and high dynamic range have come to pass, but we still prefer the general format of the superhet. Certainly, somewhere in the future, a new and better principle will evolve. But for now, let's focus our attention on the superheterodyne type of circuit. Fig 5-7 is a block diagram of a single-conversion (only one IF [intermediate frequency]) receiver. It shows the direction of the signal flow within the circuit.

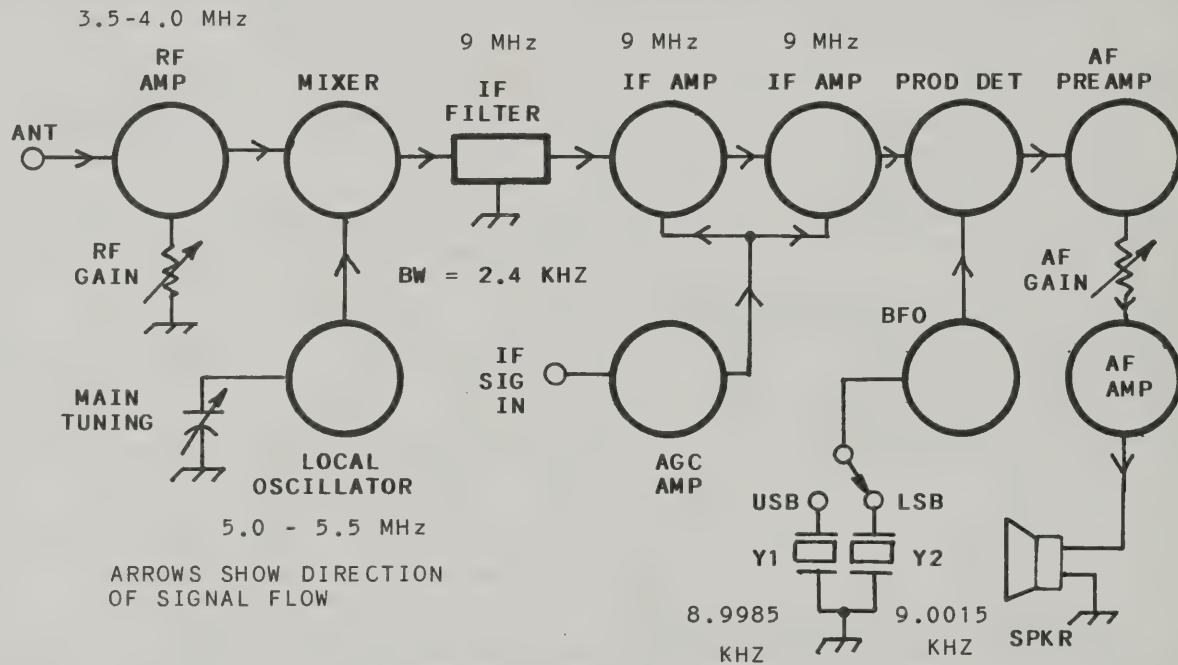


Fig 5-7 -- Block diagram of a single-conversion superheterodyne receiver. The operating frequency is 3.5 to 4.0 MHz. The IF signal is sampled at the output of the second IF amplifier and routed to the AGC amplifier. Audio-derived AGC (see Fig 5-6) may also be used if the audio energy is sampled at the output of the audio preamplifier stage.

Please refer to Fig 5-7. The signal from the antenna is amplified (10 to 20 dB) by the RF amplifier stage, which is sometimes called a "preselector." The amplified RF energy is fed to the mixer along with local-oscillator RF voltage. Output from the mixer may be taken at one of two intermediate frequencies (signal frequency plus the LO frequency, or the signal frequency minus the LO frequency). In this example we have added the signal frequency to that of the LO frequency to provide a 9-MHz intermediate frequency (IF). This is known as the "sum" frequency. With this arrangement reception at 4.0 MHz occurs when the LO is tuned to 5 MHz, and 3.5-MHz reception corresponds to an LO frequency of 5.5 MHz. Selectivity is established by means of the 9-MHz IF filter, which has a narrow bandwidth of 2.4 kHz in this example. Filters for CW reception are often as narrow in response as 250 Hz.

IF energy from the filter is amplified by the two IF amplifiers by as much as 40 dB per stage. A small amount of the 9-MHz output signal from the last IF amplifier is sampled, amplified and rectified to provide an AGC control voltage for the two IF amplifiers (and sometimes also for the RF-amplifier stage). Maximum IF gain occurs when weak signals are present, and minimum IF gain takes place when strong signals are present.

IF energy is supplied to the product detector along with beat-frequency oscillator (BFO) voltage to produce audio-frequency output from the product detector (PD). During CW reception the product of the IF and BFO signals provides a beat note that is heard in the speaker. During SSB reception the BFO supplies the missing carrier for the SSB signal (carrier reinsertion). The BFO crystals, Y1 and Y2, have frequencies that fall 1.5 kHz either side of the IF filter center frequency (9 MHz) to provide either USB or LSB reception. An audio preamplifier and an audio power amplifier are used between the PD and the speaker or earphones.

Various frequency combinations are used for the LO and IF systems. Intermediate frequencies from 455 kHz to 30 MHz are not uncommon. We must exercise care in choosing these frequencies in order to avoid frequency combinations that produce spurious responses (harmonics and mixing products) in the tuning range of the receiver.

Multiple-Conversion Receivers

A number of commercial designers use double- and triple-conversion superheterodyne circuits to enhance overall performance. The circuits are similar to that in Fig 5-7 except for additional local oscillators, mixers and intermediate frequencies. Another mixer and LO is needed for each additional frequency conversion. I prefer the single-conversion concept because it minimizes the possibility of unwanted spurious responses (called "birdies") and makes the overall receiver gain distribution easier to manage. Proper gain distribution and LO injection levels are essential to good receiver performance. Too little stage gain can degrade the receiver overall noise figure (NF), whereas too much gain at various points in the circuit can overload succeeding stages and destroy the receiver dynamic range. This subject is treated in detail in **Solid State Design for the Radio Amateur**. In simple language, the noise figure of a particular stage should be lower than that of the succeeding stage, and it should have no more gain than is absolutely necessary to override the noise of the following stage. Some amateurs build or buy external preamplifiers for use ahead of their receivers. This is seldom necessary, and more often than not the receiver dynamic range is seriously impaired by the presence of the outboard preamplifier. S-meter readings are always higher when these preamps are used, but the noise floor increases proportionally. This usually means that there is no improvement in the overall signal-to-noise (SNR) ratio of the receiver. The additional front-end gain can cause the RF amplifier and mixer stages to produce unwanted IMD products, and these may be heard in the receiver tuning range as spurious signals.

IF Filters

We experimenters like to avoid the high-cost of commercial crystal-lattice and mechanical IF filters. These cost upwards of \$50 today. There are ways to build relatively inexpensive home-made IF filters from readily available components. Fig 5-8 shows two approaches that I have used for constructing my own filters.

In the first example we can cascade some inexpensive 455-kHz IF transformers from discarded transistor AM radios. Light coupling is used between the transformers to preserve the Q and prevent unwanted bumps and dips (ripple) in across the nose of the filter response. The insertion loss of the filter is compensated by the addition of an extra IF amplifier.

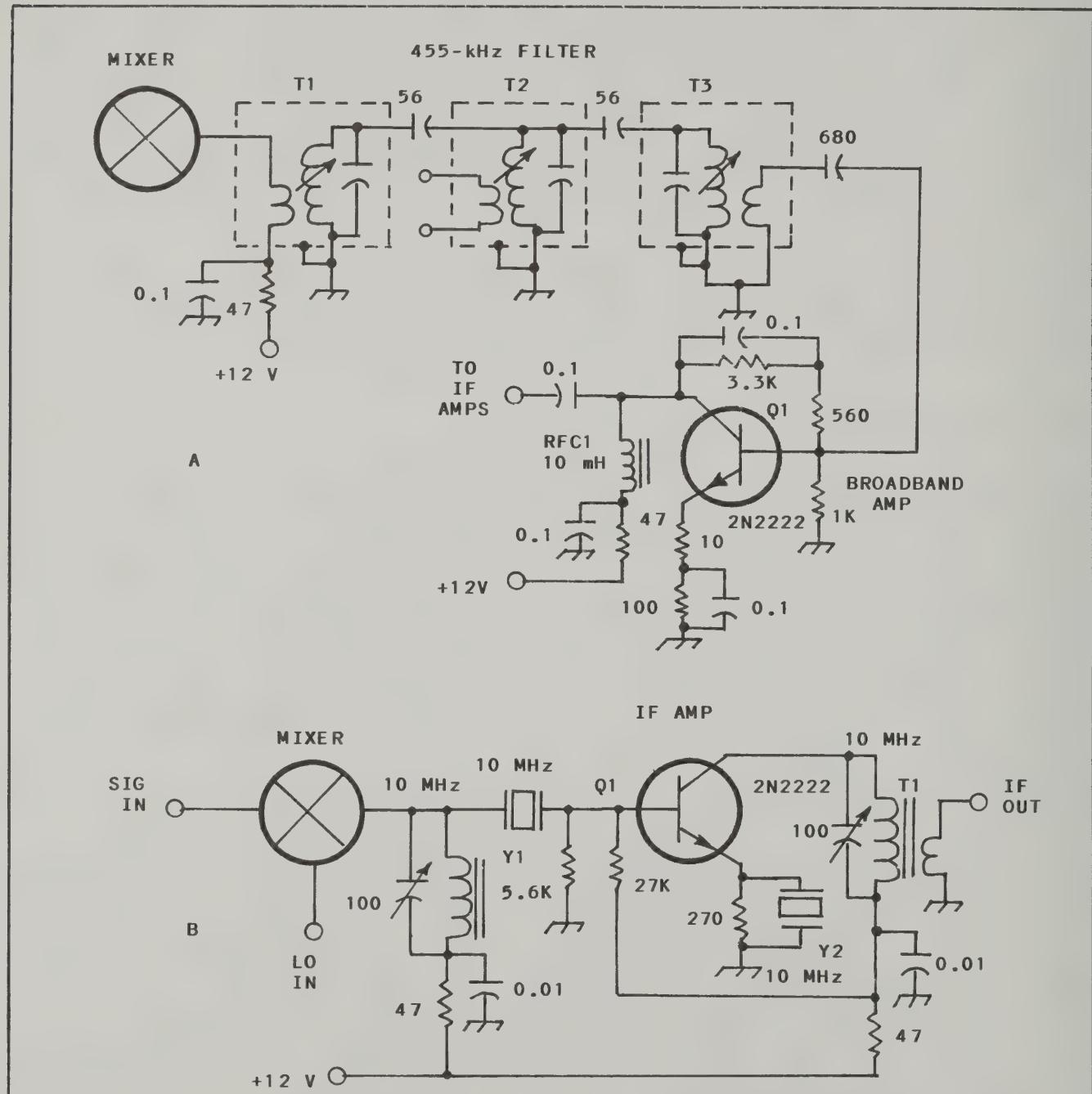


Fig 5-8 -- Circuit A shows 455-kHz IF transformers connected as an LC IF filter that is suitable for AM or SSB reception. Circuit B utilizes two-low-cost computer crystals to provide IF selectivity (see text).

Fig 5-8A shows how one or more standard transistor-radio IF transformers may be used to form a band-pass filter. The 56-pF top-coupling capacitors are selected to provide a minimum filter ripple (see earlier text). You may want to use trimmer capacitors (10-100 pF) at these circuit points to permit adjustment of the top coupling. The 3-dB bandwidth of this type of filter is on the order of 3-6 kHz, depending upon the input and output termination value and the degree of top coupling used. The unloaded Q (Q_u) of the IF transformers affects the bandwidth also. Some transformers exhibit higher Q than others. Q_1 yields some 15 dB of gain to compensate for the insertion loss of the filter. The 100-ohm emitter resistor may be replaced by a 1000-ohm trimmer control to permit adjusting the stage gain.

The Fig 5-8A scheme does not provide narrow bandwidth when using, for example, standard 10.7-MHz IF transformers. The higher the operating frequency the greater the tuned-circuit bandwidth for a specified value of Q. The tuned-circuit bandwidth doubles approximately each time the frequency is raised one octave. In other words, if the Fig 5-8A filter has a 5-kHz bandwidth at 455 kHz, it becomes 10 kHz at 910 kHz (for a specified Q), and so on.

Low-cost surplus computer crystals are used in Fig 5-8B to provide IF selectivity. Y_1 is in series with the IF signal, thereby allowing only the 10-MHz energy to pass through it. Y_2 functions as an emitter-bypass element, but only at 10 MHz. Therefore, Q_1 has very low gain at frequencies above and below 10 MHz, but yields high gain at 10 MHz. Monolithic ceramic crystals are often used in this manner by manufacturers of entertainment radios. Once again we must be concerned with bandwidth versus frequency. Low-frequency crystals have very high Q and may create bandwidths too narrow for SSB reception. I find that crystals from, say, 4 to 15 MHz are suitable for the circuit in Fig 5-8B, respective to AM or SSB reception. I use crystals that are cut for 400 to 1000 kHz for CW reception. Crystals in the 10-MHz region are nice for the Fig 5-8B circuit because we can use standard 10.7-MHz IF transformers for the mixer tuned circuit and T_1 .

I want to stress that neither circuit in Fig 5-8 is outstanding in terms of selectivity and uniformity of the band-pass response. These circuits are, however, entirely adequate for use in simple, low-cost receivers.

Ladder Filters

A high-quality crystal filter can be fashioned from inexpensive surplus crystals by using the crystals in a ladder network. An example of this type of filter is seen in Fig 5-9.

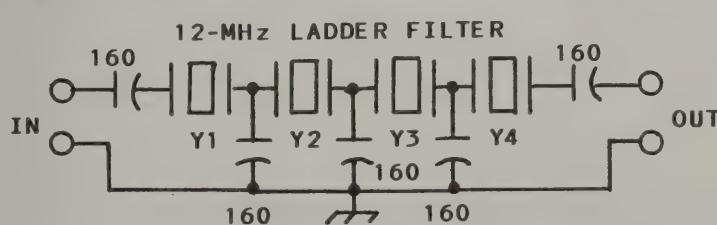
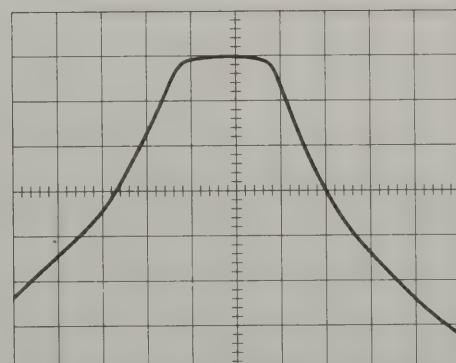


Fig 5-9 -- Practical SSB ladder filter with a 2.1-kHz bandwidth. Horiz div. at right are 1 kHz. Vert div. are each 10 dB.



The capacitor values in a ladder filter depend upon the crystal Q and series resistance. These parameters can be measured with the circuits described by W. Hayward, W7ZOI, in "A Unified Approach to the Design of Crystal Ladder Filters" (*QST* for May, 1982). Also see D. DeMaw, "A Tester for Crystal F, Q and R" (*QST* for Jan. 1990). The filter end resistance depends also upon the filter bandwidth, plus the crystal Q and R. An empirical approach to the design of ladder filters is described by Hayward in *QST* for July 1987. (See appendix.)

Ladder filters can be designed for very narrow CW bandwidths as well as for various SSB bandwidths. It is important that the crystal frequencies be closely matched. Likewise for the shunt and end capacitors. You can match the filter end impedances to the load impedances by means of broadband transformers of the appropriate turns ratio.

The filter in Fig 5-9 has an end resistance of 128 ohms. A broadband transformer with a turns ratio of 1.6:1 will match the filter to a 50-ohm termination. The capacitors are silver mica to ensure high Q and stability. The cost of the Fig 5-9 filter (using surplus crystals that cost \$1 each) is \$4. This is far better than paying upwards of \$50 for an equivalent commercial filter! I want to urge you to read the above-referenced articles. They will make the job of designing your own ladder filters an easy one.

The Local Oscillator

Frequency stability is probably the most difficult challenge you will face as an experimenter. Nobody enjoys using a receiver that won't "stay put" when it is tuned to a particular frequency. There are many causes of tunable-oscillator drift. Notable among them is internal and external heating of the frequency-determining components. RF currents flowing through the components cause them to have temperature increases, and these elevating temperatures cause changes in the internal capacitance and resistance of the components. In a like manner, variations in air temperature around the components causes value changes. Coils, capacitors, magnetic core material and the transistors themselves can undergo value changes from heating.

Another common cause of LC oscillator instability is stray RF that enters the circuit on the dc and signal lines. Energy of this type may originate from an oscillator or RF power amplifier elsewhere in the overall circuit. This kind of instability generally manifests itself in the form of abrupt frequency changes. It is for this reason that we should always build our LC oscillators in their own shield enclosures. The operating-voltage to the circuit should be filtered for RF energy by using an RF choke and bypass capacitors where the power lead enters the box. Small coaxial cable, such as RG-174, may be used for carrying the oscillator RF output voltage to its destination elsewhere in the receiver or transmitter. This will help prevent unwanted RF energy from entering the oscillator circuit. The oscillator shield box should be grounded to the equipment main frame at several points.

Zero-temperature capacitors (NPO) are best for the critical points in the VFO. An acceptable substitute is the polystyrene capacitor for frequencies up to roughly 10 MHz. These capacitors exhibit a slight negative drift characteristic, and when used in circuits that have slug-tuned or toroidal VFO coils they will help compensate for the positive drift of the core material (permeability tends to increase with heat). Silver-mica capacitors may be suitable if you grade them out for temperature stability. They are, however, the least stable of the

types under discussion here.

The effects of circulating RF current (internal heating) can be minimized by using two or more capacitors in parallel in an oscillator. For example, we may use three 33-pF NPO capacitors in parallel in place of a single 100-pF capacitor. The increased internal surface area distributes the RF current over a wider area, and hence less heating within any one capacitor.

Try to use the largest wire gauge practicable when winding your VFO coils. The larger the wire diameter the higher the coil Q (desirable). The heavier wire gauge minimizes RF heating of the winding, and this helps stability. The coil turns need to be cemented in place with a high-dielectric and stable cement such as General Cement Polystyrene Q Dope. I have made my own Q Dope by dissolving small pieces of polystyrene in carbon tetrachloride. **Caution:** Do not inhale the fumes or allow the carbon tet to contact your skin. It is considered a hazardous chemical.

If you have a VFO that uses a slug-tuned coil in the frequency-determining part of the circuit, you may discover that abrupt or long-term frequency drift occurs. This is usually caused by movement of the coil slug from vibration or changes in temperature. This malady can often be cured by melting a drop of bee's wax or canning wax on the end of the slug. This will lock it into position, but you will be able to move it later on when adjustment is needed. Do not use glue!

VFO Active Devices

Short- and long-term oscillator drift is related also to internal heating of the transistor or IC you use in your VFO. The dc and RF current within the transistor causes changes in junction capacitance, and this leads to drift. The lower the oscillator voltage the less heating. It is an easy and inexpensive matter to build up the LO-chain power with buffers or amplifiers. Regulated voltage should, of course, be applied to any LC oscillator.

Some experimenters use VVC (varactor) diodes in place of air variable tuning capacitors. Although this practice leads to lower cost and allows the circuit to be miniaturized, the VVC diode increases the frequency drift. This is because the diode junction capacitance changes when current flows through it. Again, we have both dc and RF current to consider. Therefore, I do not recommend a VVC-tuned oscillator if you are designing for optimum stability. These devices are, however, quite acceptable for medium-performance receivers and transmitters. The point I am attempting to make here is that each active device (one that requires an operating voltage) we add to an oscillator will contribute to the drift problem.

VFO Variable Capacitors

Trimmers, padders and main-tuning variable capacitors in LC oscillators should be mechanically stable. NPO trimmers help to ensure frequency stability. Small air-dielectric trimmers are best. Your main-tuning capacitor should have a bearing at each end of the rotor and it needs to rotate freely (not lumpy). Variable capacitors that have ball bearings at each end of the rotor are excellent for VFO tuning. A no-backlash vernier drive is recommended for rotating the main-tuning capacitor. Plated brass capacitor vanes are much more stable than aluminum ones. Try to avoid using aluminum-plate variable capacitors.

VFO Tuned-Circuit Q

A high-Q tuned circuit in an oscillator is mandatory for two reasons in particular: (1) The higher the Q the lower the noise at the oscillator output. (2) The higher the Q the more readily the circuit will oscillate and the more uniform the output power across the VFO tuning range.

The circuit Q can be preserved by using the lightest coupling possible between the oscillator transistor or IC and the tuned circuit. This minimizes loading on the tuned circuit, and loading lowers the Q. Use the lowest value of capacitance possible for the coupling capacitor, consistent with oscillator rapid starting and uniform output power. The Q is dependent in part on the type of core material used if you employ a toroid or slug-tuned form. No. 6 powdered-iron core material is the best for frequencies up to 50 MHz. This core material is called carbonyl J. The color code for toroids is yellow and for most slugs it is green. This mixture is the most temperature-stable of the various HF core recipes. When using a slug-tuned coil, try to set up the circuit constants so that the coil slug just enters one end of the coil. This will permit inductance adjustments and will result in minimum core-change drift. The farther the slug is moved into the coil the worse the effects of permeability changes with variations in temperature. Once a slug-tuned coil is adjusted, the tuning screw should be locked into position with a jam nut. Keep your VFO coils well away from nearby conducting objects such as the walls of the shield box. Spacing should be no less than one coil diameter from these conductors. This will also help to preserve the coil Q.

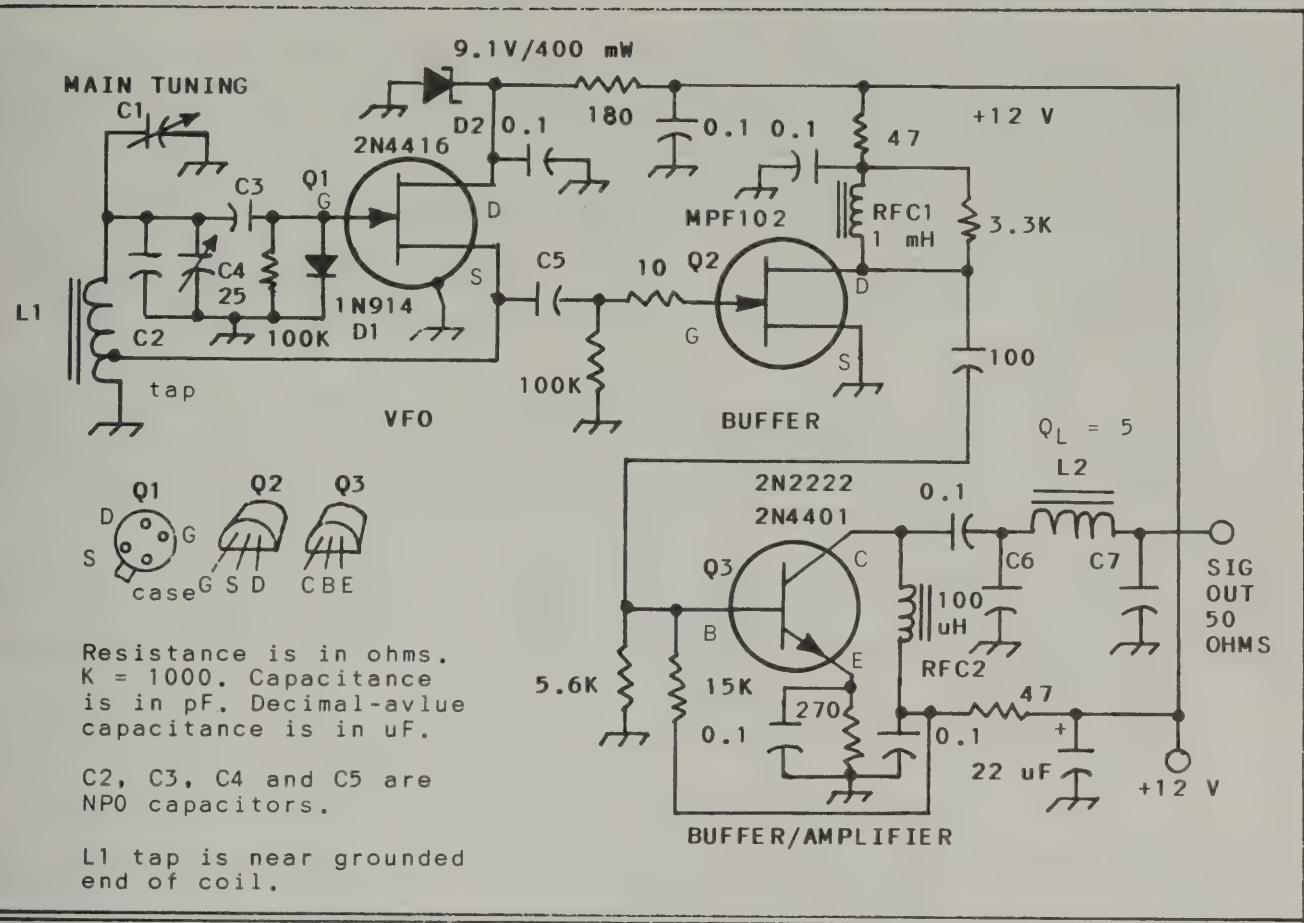
VFO Isolation or Buffering

The most stable of LC oscillators will perform poorly if they are over-coupled to their loads (subsequent stages). Excessive coupling can make an oscillator sluggish, or it may fail to operate when it is turned on. At least two buffer-amplifier stages should follow an oscillator. The coupling capacitor between the oscillator output and the input of the first buffer should have a very low value (100 pF at 80 meters, 50 pF at 40 meters, and so on). Smaller values of capacitance are even better, but the lower the capacitor value the lower the output power of the overall VFO system.

Avoid using a series resistor in combination with an output-coupling capacitor. Although this practice helps to isolate an oscillator from its load, the current that flows through the resistor generates noise that will appear in the output of the system.

VFO buffer stages isolate the oscillator from the circuit to which the VFO chain is connected. Load-level changes often occur, and these changes cause phase shifts that may affect the frequency of oscillator operation. The buffers tend to keep these load changes from being "seen" by the oscillator. See chapter 2 (Fig 2-15) for an example of a practical VFO that has two buffer stages. The first buffer is a JFET. This transistor has a very high input impedance (100 K ohms) that is set by the gate resistor. This high-impedance termination for the oscillator results in minimal loading. A bipolar transistor is used as the second buffer. It amplifies the signal and provides additional VFO isolation.

Fig 5-10 contains a practical circuit for a "universal" VFO. You may use it as a basis for designing a VFO for your receivers or transmitters.



f(MHz)	C1	C3	C2, C5	C6	C7	L1	L2
1.8-2	100	220	100	400	910	24 uH, 71t no. 30 on T68-6 toroid	60t no. 30 on T50-2 toroid, 17.8 uH.
3.5-4	50	150	68	200	470	9.5 uH, 44t no. 26 on a T68-6 toroid.	9 uH, 47t no. 26 on a T50-6 toroid.
5-5.5	50	130	50	150	330	5 uH, 33t no. 24 On T68-6 toroid.	6.4 uH, 40t no. 28 on a T50-6 toroid.
7-7.3	25	110	50	100	240	3.6 uH, 27t no. 24 on a T68-6 toroid.	4.7 uH, 35t no. 26 on a T50-6 toroid.

Fig 5-10—Circuit of the universal VFO. Fixed-value capacitors in the chart above are NPO ceramic. C1 is an air variable. Resistors are 1/4-W carbon film or carbon composition. The Q1 case is grounded. If Q3 should tend to self-oscillate, place a 3.3K-ohm resistor in parallel with L2. Toroid cores are available from Amidon Assoc., Inc., PO Box 956, Torrance, CA 90508. VFO tuning range for each band extends above and below the listed frequencies. The L1 tap is 1/4 the total coil turns near the grounded end of L1 (11 turns for 80 meters, and so on).

The Fig 5-10 VFO is structured for specified operating-frequency ranges. You may alter the circuit constants for various tuning ranges that work into the scheme of your receiver design. The frequencies and component values listed in the figure can serve as approximations when tailoring the circuit for other frequencies. A rough rule of thumb suggests determining the reactance values for the various critical capacitors and inductors, then using the reactance values for calculating the approximate values for other frequency ranges. Equations for learning the XC and XL reactance values are presented in **The ARRL Electronics Data Book** and **The ARRL Handbook**. They involve only simple algebra and are no more complicated than the Ohm's Law formulas.

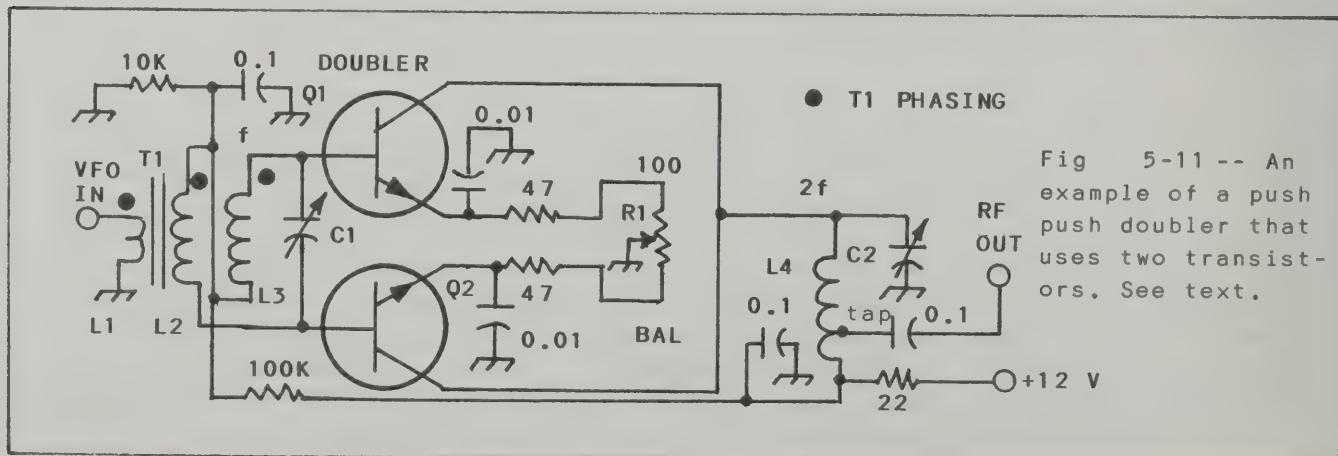
Do not use double-sided PC board for any VFO circuit. The copper conductors on the etched side of the board act as capacitor plates in combination with the ground-plane side of the board. The glass-epoxy insulation between them is a poor dielectric at HF, and these unwanted capacitors are extremely unstable.

Keep all of the signal leads short and direct in any VFO. Similarly, the leads on the capacitors and resistors need to be short to prevent unwanted stray inductance. Make certain the capacitors and resistors are snugly against the component side of the PC board.

There is no reason why you can't use a quality slug-tuned inductor for L1 in Fig 5-10. In fact, I prefer them to toroid cores because there is less magnetic core material to cause drift. Blank, ceramic slug-tuned coil forms are expensive and scarce, but don't hesitate to use them in the interest of good VFO performance. Make certain the core material is powdered iron, and that it is the no. 6 material mentioned earlier, which is specified particularly for 20-50 MHz use. C4, which was not mentioned in the caption for Fig 5-10 is a 25-pF NPO ceramic trimmer. You may also use a 25-pF miniature air-variable trimmer. Avoid using mica trimmers or those made from plastic.

Coat L1 in Fig 5-10 with two or three applications of high-Q coil cement, such as Polystyrene Q Dope. This will keep the coil turns firmly in place and will prevent moisture from changing the coil characteristics.

The universal VFO may be used for generating frequencies above 7.3 MHz by driving a push-push doubler with the Q3 output power. Fig 5-11 shows how a push-push doubler may be constructed.



You may use any HF transistor for Q1 and Q2 in Fig 5-11. I suggest the 2N2222A or 2N4400/2N4401 types of transistors for low-power applications. A power type of doubler might use 2N3866s or similar TO-5 devices. T1 is tuned to the driving or fundamental frequency (f). L4 is tuned to the second harmonic of the input frequency. L2 and L3 are a bifilar winding with sufficient inductance to permit resonance with C1 at f. A 300-pF trimmer is suitable for C1 when the input frequency is from, say, 7 to 14 MHz. a 100-pF trimmer may be used from 14 to 30 MHz. C2 is a 100-pF trimmer. L4 has the proper inductance value to cause resonance with C2 at twice the doubler input frequency.

Circuit balance, by means of PC control R1, is necessary in order to minimize the presence of the doubler input energy at the output of L4. Adjustment may be carried out while sampling the doubler output and monitoring it with a receiver that is tuned to the doubler input frequency. Set R1 for minimum S-meter indication, respective to f. A scope can be used for this adjustment by adjusting R1 for the purest waveform obtainable at the doubler output port. The L4 tap is selected in accordance with the load into which the doubler operates. Maximum power transfer occurs when the tap is located correctly. Tap L4 at 10% of its total turns (near the +12-V end) for loads of 100 ohms or less. The tap will need to be moved higher up on L1 for loads that have a higher impedance. You can broaden the response of T1 and L4, if necessary, by placing a 2.7K-ohm resistor in parallel with C1 and C2. This will, however, lower the doubler output power.

A push-push doubler operates as efficiently as a straight-through RF amplifier. It must operate in class C in order to provide good efficiency. The forward bias on Q1 and Q2 in Fig 5-11 makes the doubler easier to drive, but does not cause it to function in class A. The driving power sends the doubler into the desired class C mode. Suppression of the fundamental input signal is on the order of 40 dB if the circuit is well balanced and the Q of the output tuned circuit is reasonably high.

Single-ended doublers and triplers are not efficient (40% for a doubler and 25% for a tripler, typically, when using bipolar transistors). Push-pull triplers are better than single-ended ones, and they help to attenuate harmonics. Harmonic rejection is generally poor with single-ended frequency multipliers. Likewise with suppression of the fundamental energy that is supplied to the frequency multiplier.

VFO stability is easier to achieve, for example, at 7 MHz than it would be at 14 MHz. Therefore, the use of a doubler is helpful when it is necessary to generate LO energy at the higher frequencies.

IF Amplifiers

The overall gain of any sensitive receiver is on the order of 90-100 dB. The receiver gain distribution we discussed earlier is dependent in part on the gain of the IF amplifier section. Also, if AGC is incorporated in the design, two IF amplifiers are necessary in order to ensure good AGC dynamic range. Among the more popular IF-amplifier ICs are the Motorola MC1350P, MC1590G and RCA CA-3028A, although there is a host of other excellent chips. Generally, the Motorola and RCA components are easier to obtain in small lots than are other U.S. brands.

It is not essential that IF amplifier have tuned input and output circuits. If a narrow-bandwidth IF filter is used ahead of the 1st IF amplifier, the two IF

stages can be used as broadband amplifiers. There will be very little difference in gain either way. Tuned circuits are used mainly to establish selectivity. Stability is easier to realize when using untuned amplifiers -- a plus feature. On the "down side" of the broadband concept, wide-band noise is generally greater when we don't use narrow-band tuned IF amplifiers. Impedance matching is, however, mandatory between stages to ensure maximum available amplifier gain. This is true of broadband and narrow-band amplifiers. The more elegant superhet receivers of today use two IF filters. The narrowest one is placed ahead of the 1st IF amplifier and the second one is located just after the last IF amplifier. The second filter has a wider response than the first one, and the center frequency of each filter must be the same. The second or "tail-end" filter greatly reduces wide-band noise that is generated within the IF-amplifier chain. The net effect is an improvement in overall receiver noise figure. The tail-end filter need not have as many poles as the first filter. For example, we might use a 2.2-kHz bandwidth, 6-pole filter immediately after the mixer. The tail-end filter may have only two poles and a bandwidth of 3 kHz or greater.

Bipolar transistors or dual-gate MOSFETs may also be used as IF amplifiers. Good AGC dynamic range is generally less difficult to obtain when using ICs that are designed for the application. There is no rule that says we must use AGC in a receiver, and many hams build simple receivers that do not have AGC. Rather, they use a manual gain control for varying the gain of the IF strip. Fig 5-12 illustrates a practical IF amplifier that uses RCA CA3028A ICs.

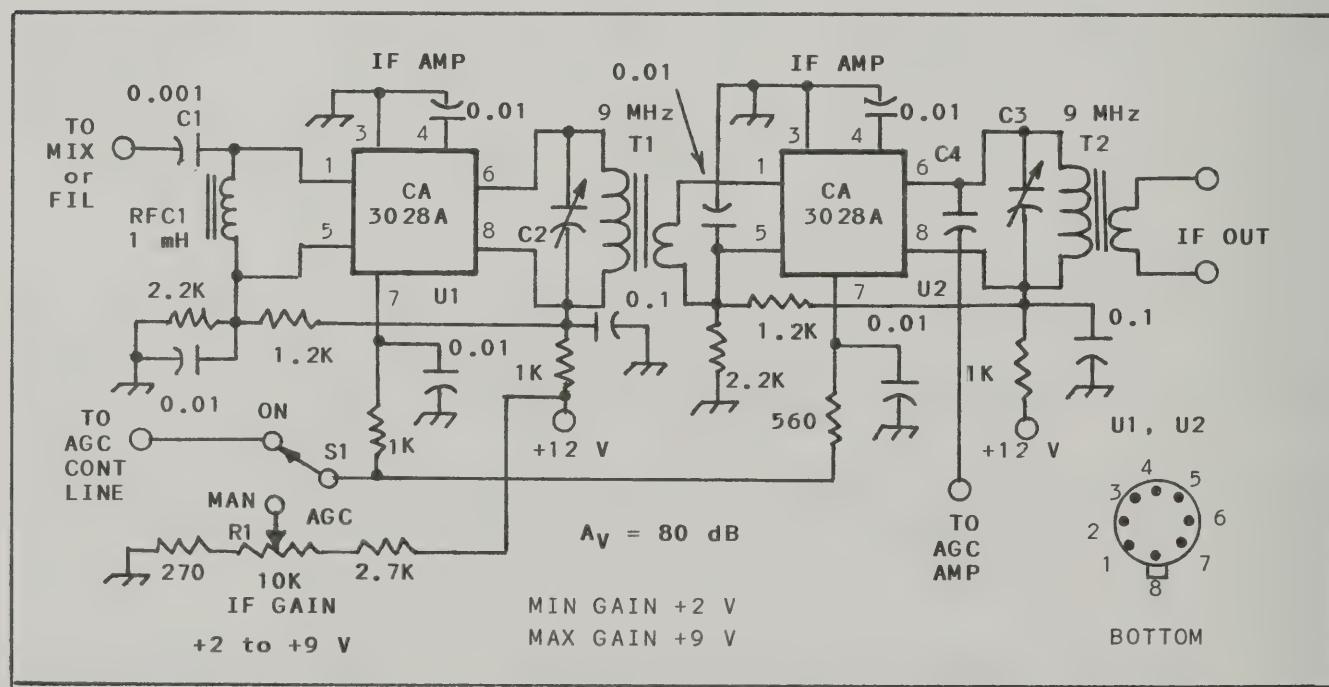


Fig 5-12 -- Schematic diagram of a practical IF amplifier. Frequencies from 455 kHz to 30 MHz may be used with CA3028A ICs. This is based on the popular 9-MHz IF. Fixed-value capacitors are disc ceramic. Resistors are 1/4-W carbon film. The C1 value is chosen in accordance with the output Z of the mixer or IF filter. C2 and C3 are 150-pF trimmers. T1 primary is 3 uH (27 t no. 28 enam. wire on a T50-2 toroid). T1 secondary has 6 turns of no. 28 wire. The T2 primary is the same as for T1. The secondary turns are determined by the load presented by the prod. det. CA3028A input Z is 1000 ohms, and output Z is 4K ohms. R1 is a linear-taper carbon control.

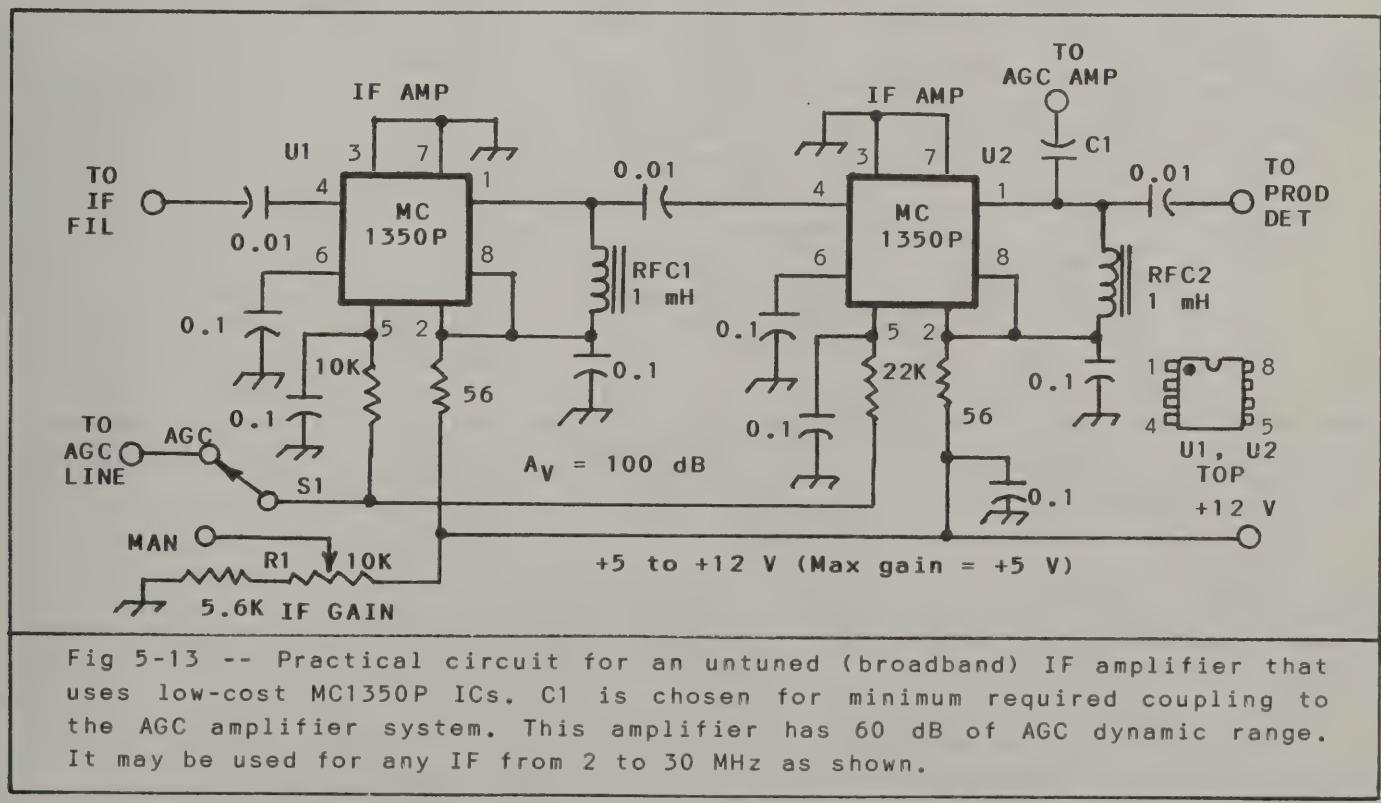
IF energy is sampled at pin 6 of U2 via C4. This capacitor should be as small in value as practicable, consistent with full AGC range. The larger you make C4 the greater its effect on the C3/T2 tuned circuit. A typical C4 value for 9 MHz is 27 pF. In an ideal situation the sampled IF energy is fed to the gate of a JFET AGC amplifier or a similar high-impedance load. This practice assures minimum interaction between the two circuits.

S1 in Fig 5-12 provides an option between AGC and manual IF gain control. R1 and its end resistors should be set up to provide a voltage range of +2 to +9 volts, since this is the prescribed range for the CA3028A. The AGC voltage should also be +2 to +9 volts.

Instability is always lurking in high-gain IF amplifiers. Therefore, it is imperative that you keep all signal leads short. Bypass capacitors need to be located at the IC pins, where applicable. Use low-profile IC sockets if you don't mount the ICs directly on the PC board. Should instability still become manifest, try placing a 3.3K-ohm resistor in parallel with the primary winding of each tuned circuit (T1 and T2). This technique is a "bandaid" of sorts, and it will broaden the response of the tuned circuits. This is not an important matter if you use a selective IF filter ahead of U1. In other words, "when all else fails, swamp the tuned circuit."

MC1350P or MC1590G IF Amplifier

Fig 5-13 contains the circuit for a practical IF amplifier strip that employs Motorola MC1350P or MC1590G ICs. Each type of IC offers the same overall performance, but the pin-out for these chips is different. The MC1350P is lower in cost than the MC1590G, It has an 8-pin DIP format and it is easy to obtain.



The Fig 5-13 circuit differs from that in Fig 5-12 in two major ways. First, the IF amplifier gain is maximum at the lowest AGC voltage when using MC1350 or MC1590G ICs. The opposite is true of the CA3028A. Secondly, the Fig 5-13 IF strip is untuned. As shown, it may be used for any IF from 2 to 30 MHz. RFC1 and RFC2 should be changed to 10-mH units for 455-kHz operation. There is no reason why the Fig 5-13 circuit can't be changed to a narrow-band amplifier, as shown in Fig 5-12. The tuned circuits are necessary if no IF filter is used ahead of U1. The CA3028A amplifier in Fig 5-12 can be converted to an untuned circuit by replacing T1 and T2 with RF chokes. Should you encounter instability with the MC1350Ps, simply put a 1.5K-ohm resistor in parallel with each of the RF chokes. This will not degrade the amplifier performance. In fact, the W7Z0I "Competition Grade Receiver" that is described in **Solid State Design for the Radio Amateur** uses carbon resistors in place of RF chokes. That IF amplifier uses MC1590Gs. The W7Z0I receiver incorporates an excellent AGC amplifier and control system that is suitable for use with the Fig 5-13 circuit.

AGC System for the CA3028A IF Amplifier

An AGC and S-meter control circuit for the Fig 5-12 circuit is depicted in Fig 5-14. It was used by W1FB in "His Eminence, the Receiver," which is also described in **Solid State Design**.

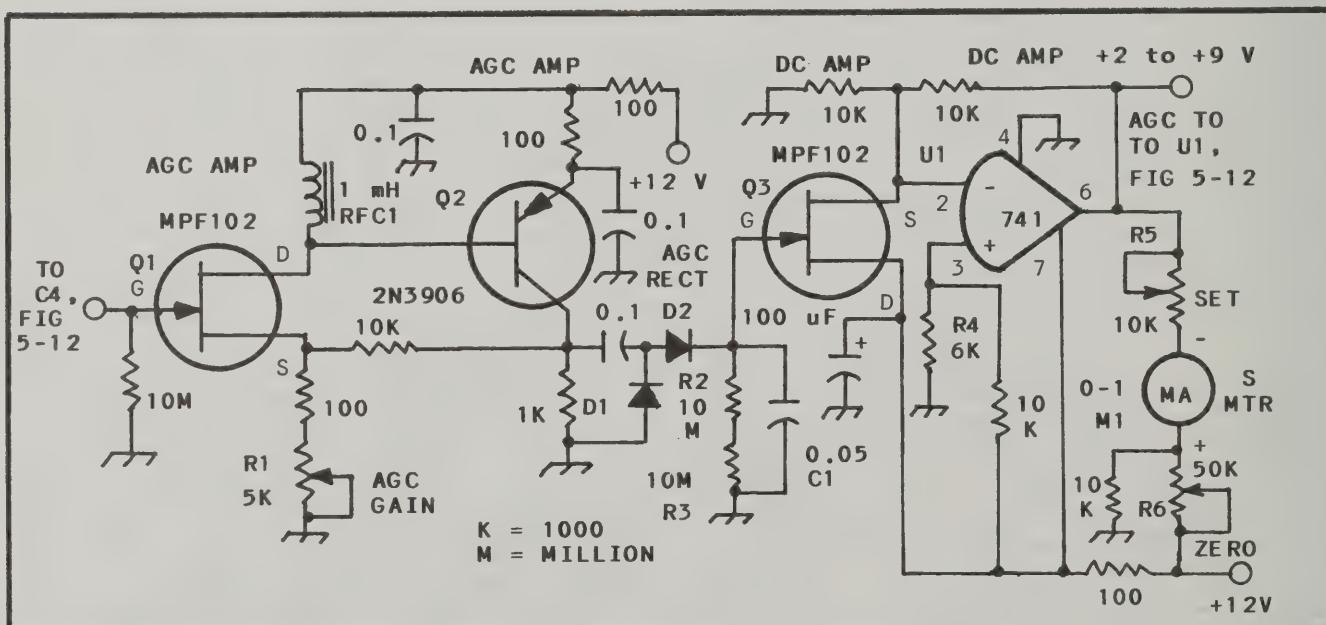


Fig 5-14 -- A practical IF-derived AGC system for CA3028A IF amplifiers. An audio-derived AGC circuit for CA3028As is shown in Fig 5-6. R1 is set for the signal level at which AGC action commences (an S2 or S3 signal should cause D1 and D2 to commence rectifying the sampled IF energy. R4 is chosen to provide +9 V at pin 6 of U1. R5 and R6 are PC-mount controls. R5 is adjusted for a full-scale S-meter reading and R6 sets the meter to zero when no signal is present. C1, R2 and R3 establish the AGC time decay time. D1 and D2 are 1N914 or equivalent silicon diodes. Capacitors are disc ceramic, 50 V or greater. Fixed-value resistors are 1/4-W carbon film. R1, R5 and R6 are carbon-composition linear PC-mount controls. Change RFC1 to 2.5 mH for IF systems below 2 MHz. The polarized 100-uF capacitor is 25 V tantalum or electrolytic.

Mixers and Product Detectors

Mixers and product detectors are described in the early chapters of this book. There is no point in covering the same subject again. Each of these circuits functions in a similar manner, and the same general rules apply in each case. It is important that you use the correct amount of LO or BFO injection. Too little or too much injection power or voltage leads to inferior mixer or detector performance. Take the time to measure the injection level while the mixer or product detector is operating. Adjust it for the prescribed value by raising or lowering the output power of the VFO or BFO.

Beat Frequency Oscillator

The beat-frequency oscillator (BFO) is as important as any other circuit in your receiver. It needs to be stable, provide the required output power, and the output signal must be free of hum and noise, within reasonable constraints.

It is possible to construct an acceptable LC type of BFO for low frequencies, such as 455 kHz. Stability is easy to achieve, and the tunable BFO allows you to move the BFO frequency about in the IF pass band. This can be advantageous when dealing with certain types of QRM. Many early day ham receivers were equipped with tunable BFOs that were adjustable from the receiver front panel.

Modern receivers have crystal-controlled BFOs. The frequency stability they provide is essential in consideration of the high orders of IF selectivity we use. Fig 5-15 contains the circuit for a practical BFO that has provisions for USB and LSB by means of switched oscillators. It has more output power than is needed for most product detectors. It's always better to have too much than too little available output power from a VFO or BFO!

The Fig 5-15 circuit has a pair of JFETs that operate as crystal oscillators. They are arranged in a standard Pierce circuit. C1, C2, C5 and C6 are feedback capacitors. The smaller the C1 and C2 value the higher the operating frequency. You may wish to replace these fixed-value capacitors with 50-pF trimmers to allow some trimming of the oscillation frequency. S1 switches the dc voltage from Q1 to Q2 for changing from USB to LSB operation.

More extensive control of the operating frequency may be had by inserting a 50-pF trimmer capacitor and a 5-uH RF choke in series between Y1 and the junction of C3 and C5. This provides VXO action, and hence a much wider range of frequency control. Larger values of inductance result in greater frequency shift below the marked value of the crystal. When using a 9-MHz crystal, for example, you should be able to shift the operating frequency 1 kHz above the Y1 marked frequency and 3 kHz below it with the values suggested. The lower the basic crystal frequency the smaller the available frequency change.

Q3 in Fig 5-15 is a JFET source follower. It has minimum loading effect on the two oscillators by virtue of the high input impedance of Q3. Output from Q3 is approximately 0.9 the input voltage, which is the rule for source followers. Output from Q3 is at low impedance ($Z = 10^6/\text{gm}$) for providing a reasonable match to the base of Q4, thus ensuring maximum power transfer.

Q4 is a class A amplifier that yields roughly 42 mW of maximum output power at 50 ohms. This equates to 1.4 V RMS or +16 dBm.

The output power for Q4 of Fig 5-15 may be set by selection of the value of the emitter resistor, R10. The greater the R10 value the lower the output power. T1 is tuned to resonance at the crystal frequency. R13 may be used as a carrier-insertion control if the BFO is used as a carrier generator in an SSB exciter. Eliminate R13 if you use this circuit for the BFO in a receiver.

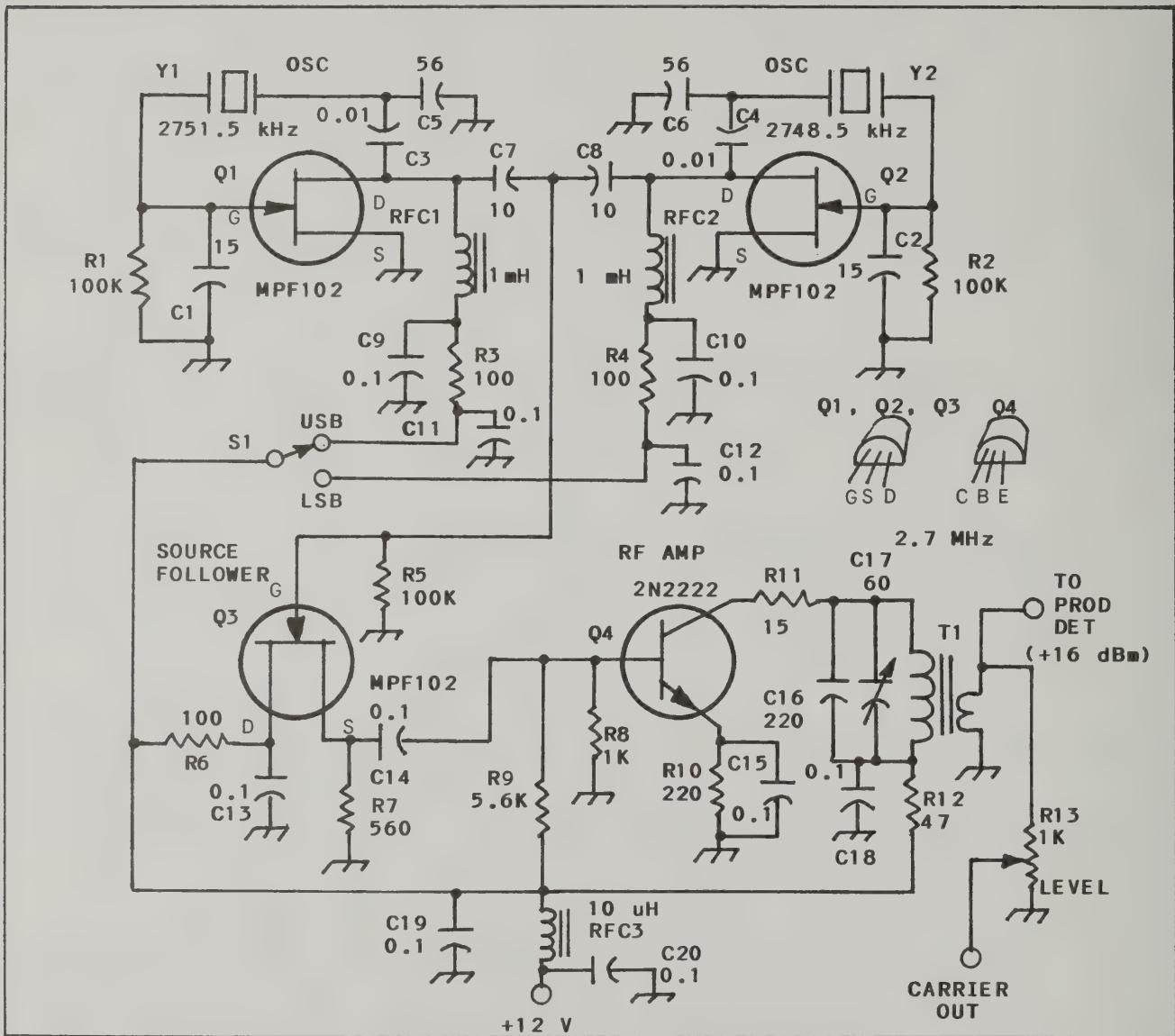


Fig 5-15 -- Circuit for a practical two-channel BFO that has more than ample output power for most applications. This circuit may be used also as a carrier generator in an SSB transmitter, with R13 serving as a carrier-insertion level control. The circuit is suitable for use on frequencies from 1 MHz to 30 MHz by changing the crystals, feedback capacitors and the output tuned circuit accordingly. Disc ceramic capacitors are used for all but C17. Fixed-value resistors are 1/4-W carbon film. R13 is a carbon-composition, linear-taper control. S1 is a SPDT rotary or toggle type, panel mounted. T1 primary inductance is 16 μ H and has 17 turns of no. 26 enam. wire on an Amidon FT-37-61 ferrite toroid. T1 link has 3 turns of no. 26 enam. wire. C17 is a 60-pF plastic trimmer.

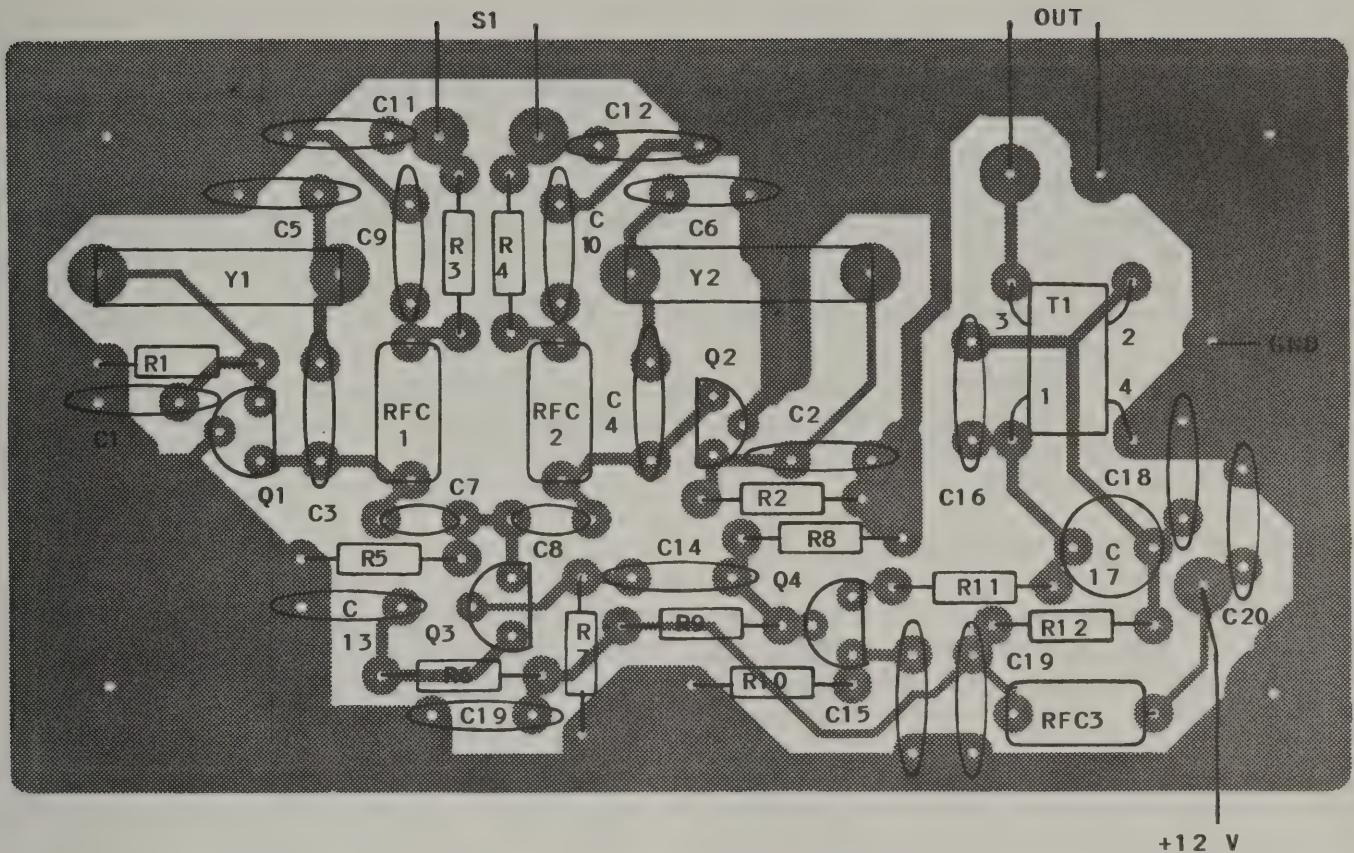
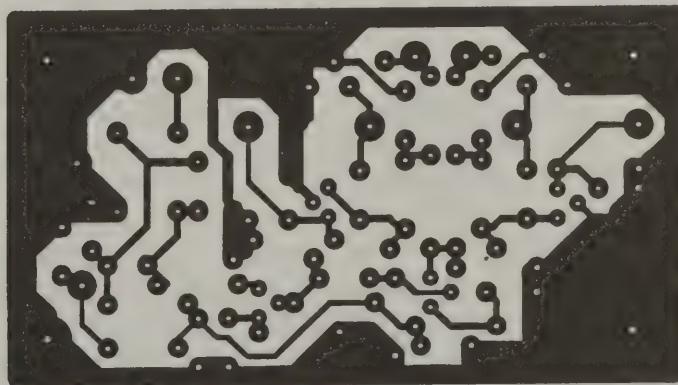


Fig 5-16 -- Xray view of the component side of the BFO PC board (not to scale). Component shapes are not necessarily representative of the actual component shapes. Etched, drilled and plated PC boards for this project are available from FAR Circuits, 18N640 Field Ct., Dundee, IL 60118.



Etched side of circuit board, shown to scale.

A parts placement guide and an etching template for the Fig 5-15 circuit are presented in Fig 5-16. Circuit boards are available (see Fig 5-16 caption).

There is no reason why the BFO can't be used also as a QRP transmitter. It should be followed, if this is done, by a 50-ohm pi-section harmonic filter to ensure minimum harmonic output. Keying can be accomplished by allowing the oscillator to run while keying the +12-V supply line to Q3 and Q4. The circuit would thus provide a two-frequency 50-mW transmitter, or it can be used to excite a class C transistor amplifier for RF power levels up to 1.5 W.

Extensive lead filtering and stage decoupling is evident when you examine Fig 5-15. This was done to keep BFO energy from reaching only the desired injection point at the product detector. My version of this circuit is contained in a shield box made from PC-board sections.

Filter Switching with Diodes

It is awkward to switch RF front-end and IF filters by means of a mechanical switch. We may use diodes for this purpose in the interest of eliminating long signal leads that can easily degrade the filter performance. The diodes are turned on and off for selecting the desired filter by way of a mechanical switch that routes a dc control voltage to the switching diodes. This is illustrated in Fig 5-17.

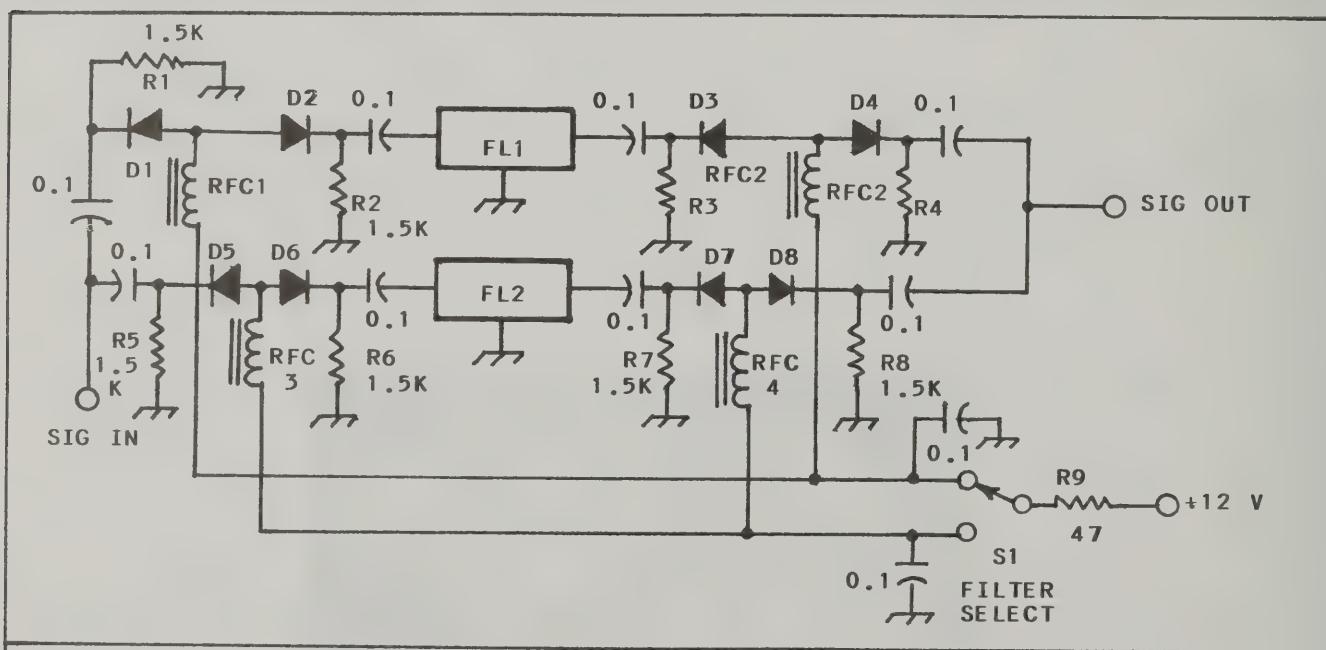


Fig 5-17 -- Example of diode-switched RF or IF filters. D1-D8, inc., are type 1N914. Capacitance is in μ F. Resistance is in ohms. Resistors are 1/4-W carbon film or carbon composition. RF chokes are mini 10-mH for frequencies up to 1 MHz. Use 1 mH for frequencies from 1 MHz to 30 MHz. S1 is a DPST toggle or wafer type. This circuit is suitable for filters with terminal impedances of 100 ohms or less. See text for a discussion of parts values and methods when FL1 and FL2 have impedances greater than 100 ohms.

Care must be taken to avoid terminating any switched filter with too low a resistance. If you use resistors as they are applied in Fig 5-17 be sure that they are at least 10 times greater in value than the characteristic impedance of the filter, since the ends of the filters and the diode-biasing resistors are effectively in parallel. The Fig 5-17 circuit is for switched filters that have 50-ohm impedances. RF bandpass filters with 50-ohm input and output coupling links are examples of filters that may be used for FL1 and FL2.

R1 through R8 in Fig 5-17 should be no greater in value than 1.5K ohms. Larger values will prevent the diodes from hard-switching to the ON state. Diodes that are not turned on completely are fairly resistive, and when using them in series with the signal path there will be significant losses. If the Fig 5-17 resistors were changed to, say, 2.7K ohms, the insertion loss through the system would be on the order of 6 dB or greater.

Suppose that you are using IF filters that have a 2000-ohm characteristic impedance. What can you do to the Fig 5-17 circuit to allow the use of these filters? R1 through R8 need to be replaced by RF chokes of the same value as RFC1 and RFC2. We now have only 47 ohms of resistance (R9) in the diode supply line. This would cause the diodes to fail from excessive current. This can be remedied by changing R9 to a 1K-ohm resistor.

You will note that back-to-back diodes are located at each switching point. A single diode could be used at those circuit locations, but poor isolation of the filters would result. The problem becomes worse as the filter frequency is increased (leakage). Therefore, I recommend that you use two diodes at each solid-state switch location, as shown in Fig 5-17. You may switch as many filters as required by simply adding additional diodes, filters, capacitors, RF chokes and resistors, as illustrated. S1 will then require additional terminals in order to accommodate the extra filters. Examples of applications for which you might use the Fig 5-17 circuit are when switching from an SSB to a CW IF filter in a receiver, or when switching bandpass filters in a receiver front end. The diodes are too fragile to be used in filters for RF power amplifiers.

Four-Pole RC Active CW Filter

In the absence of selective IF filters for CW reception we may use an active audio filter of the RC (resistance-capacitance) type to provide a peak audio response at some chosen frequency. Most CW operators prefer a CW beat note that occurs between 500 and 800 Hz. The filter shown in Fig 5-18 has provisions for switching in as many as four poles. As additional poles or filter sections are added, we enjoy the benefits of sharper skirt response (improved selectivity). In fact, with all four poles activated you may observe an annoying "ringing" (an echo or pinging type of sound on the CW beat note) effect, but a weak CW signal will stand out above the background noise when using this degree of selectivity. Normally, we will use one or two filter sections for average and good signal levels.

Although the Fig 5-18 filter may be used outboard from the receiver at the headphone jack, it is better to connect it between the audio preamplifier and the audio-output stage of the receiver. This helps prevent excessive audio power from overloading the filter.

The audio filter in Fig 5-18 has a voltage gain of 2.1 and a Q of 3. Therefore, it will boost the audio signal approximately 6.5 dB. The peak frequency for

the filter is 750 Hz. You may use two inexpensive uA747 dual op amps in this circuit, but a quieter filter (less internal noise generation) may be had by using TL082 dual bi-FET op amps. The pinout is the same for either type of IC.

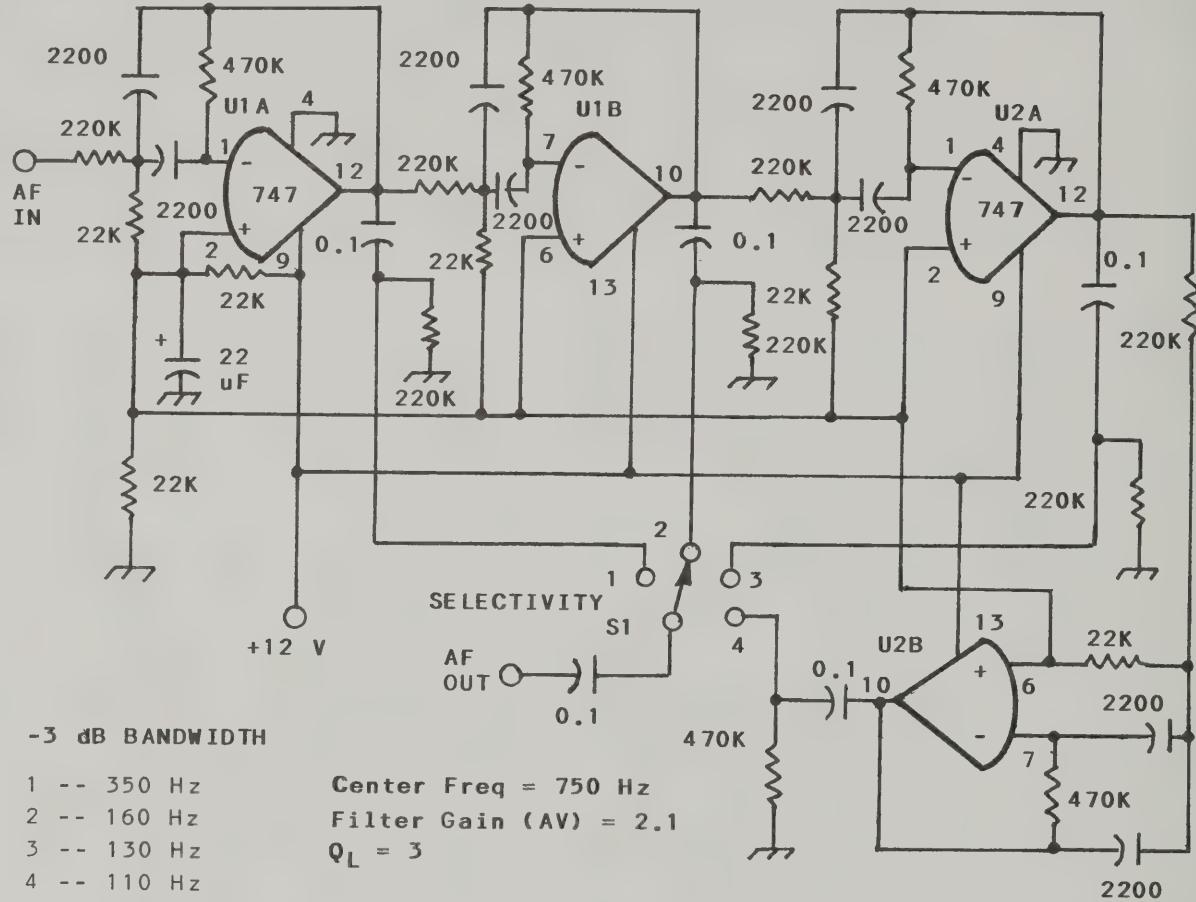


Fig 5-18 -- Practical circuit for a four-pole RC active audio filter that is suitable for CW reception. Resistors are 1/4-W carbon film or carbon composition. The resistors should be matched for value within 5%. Decimal-value capacitors are in μF and are disc ceramic. The polarized capacitor is electrolytic or tantalum, 16 V or greater. The frequency-determining capacitors are in pF (2200 pF) and need to be matched within 5%. Polystyrene capacitors are recommended. S1 is a single-pole, 4-position phenolic or ceramic single-wafer switch. S1 may be replaced with a switch that has more poles and positions to permit bypassing the filter for SSB reception.

It is practical to have provisions for removing the audio filter from the receiver audio system when you do not wish to operate the CW mode. S1 of Fig 5-18 may be changed to a three-pole, five-position switch for this purpose. One switch section can be used to turn the operating voltage on and off. Another switch section can be utilized at the AF input port of the filter to allow the audio to be routed around the filter when you are copying SSB or some other wide-band signal. An extra position on the third wafer (S1 of Fig 5-18) may be used to complete the original receiver audio circuit during filter non-use. If, on the other hand, you plan to use the filter in a CW-only receiver you may use the circuit as it is shown above.

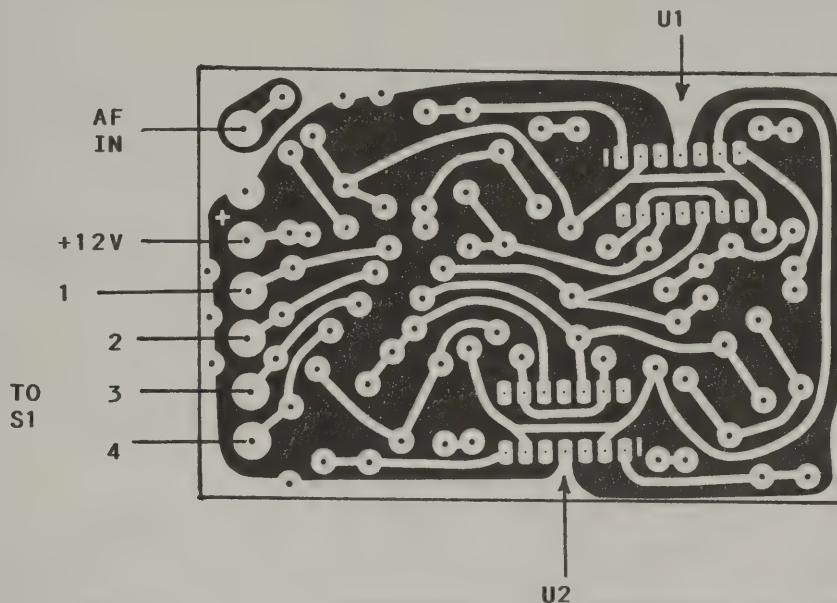


Fig 5-19 -- Scale pattern for the audio filter in Fig 5-18 as seen from the etched side of the PC board. The black areas are etched away and the light areas are copper conductors. Pin no. 1 of each IC is identified, as are some other key points that may be used to determine parts placement. A parts-placement overlay for this project was not available at the time this book was written.

A 40-dB Gain Preamp for 160 M

When we use small receiving loops and Beverage antennas for receiving it is often necessary to use a low-noise preamplifier between the antenna and the input of the station receiver. Ideally, the preamplifier provides the necessary gain to make the incoming weak signal on par with the same signal when it is received via the transmitting antenna. A preamplifier with an RF gain control makes this possible. The input device of the preamp needs to have a low noise figure in order to avoid amplifying the internal noise of the first stage in the preamp. This is of special significance when dealing with very weak signal levels from the receiving antenna. A small receiving loop may deliver a signal that is 25 dB or greater below the strength of the same signal when it is received by a full-size antenna. The same is generally true of a Beverage antenna. Too little signal strength makes it necessary to operate the receiver with its audio gain wide open, or nearly so, in order to have ample speaker or headphone volume. A high-gain preamp eliminates this problem.

Fig 5-20 contains a circuit for a practical preamp that has variable RF gain. Q1 is a grounded gate JFET low-noise amplifier. It has a gain of roughly 10 dB. A broadband input transformer (T1) transforms the 50-ohm antenna impedance to the 200-ohm input impedance of Q1. D1 and D2 are protection diodes that conduct when the energy on the antenna exceeds 0.7 V peak. This protects the preamp from transmitted energy that may appear on the receiving antenna.

A tuned circuit is used at the output of Q1. RF energy from Q1 is fed to U1, which is an IF-amplifier IC that can provide up to 40 dB of gain. R1 sets the gain of U1 to the desired level. Output from U1 is supplied to a broadband step-down transformer which provides an impedance match to a 50-ohm receiver input

port. A printed-circuit board and parts-placement guide for this circuit (an 80/160-meter version) are available from FAR Circuits (see page 106 for address).

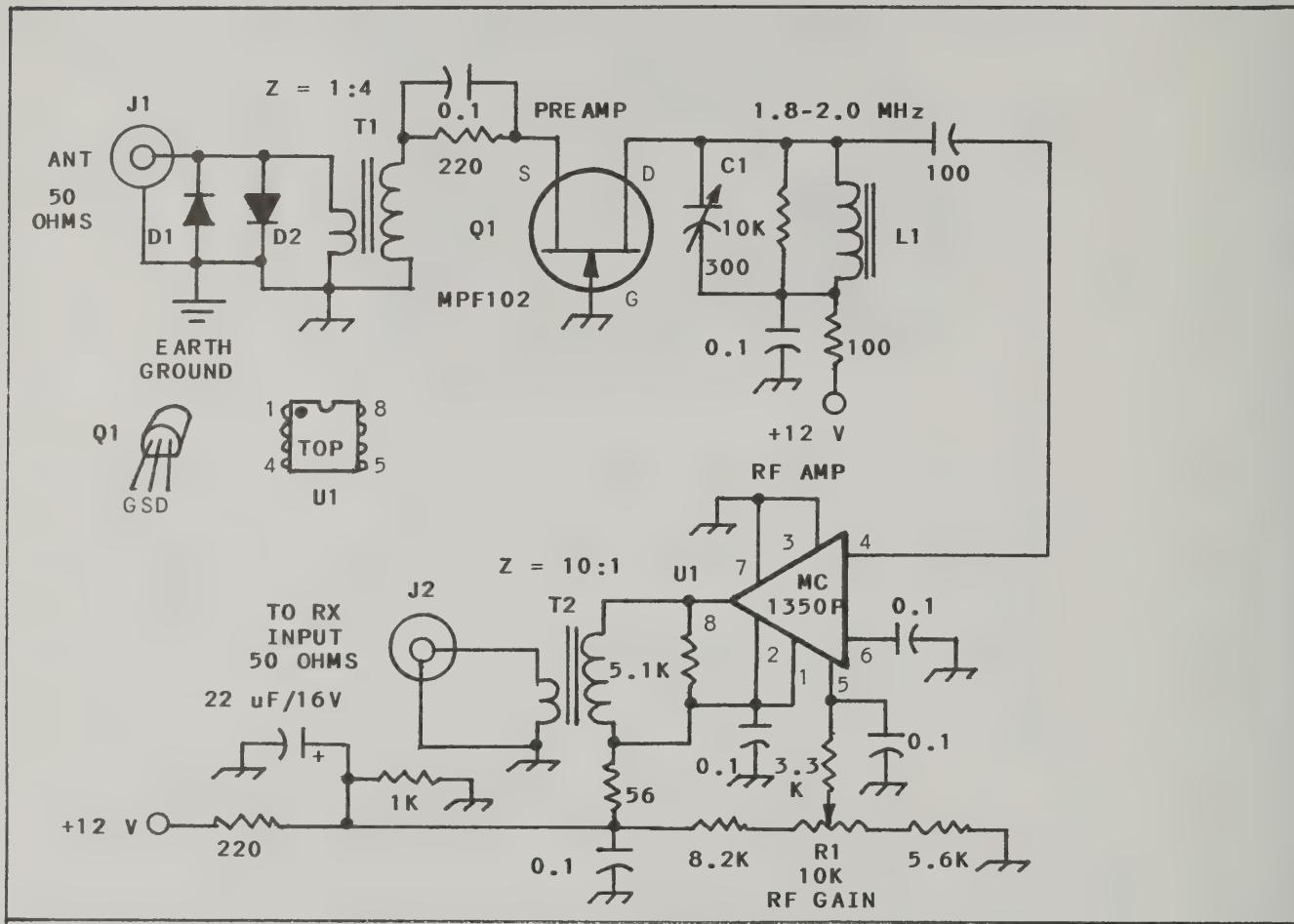


Fig 5-20 -- Schematic diagram of a 40-dB preamplifier for 160-meter Beverage or small loop antennas. Fixed value capacitors are disc ceramic. The polarized capacitor is tantalum or electrolytic. Resistors are 1/4-W carbon film or carbon composition. Circuit may be used for other bands by changing the inductance value of L1. C1 is a 300-pF mica trimmer (ARCO no. 427). D1 and D2 are type 1N914 silicon diodes. L1 = 32 μ H (22 turns of no. 26 enam. wire on an Amidon FT-50-61 (125 μ u) ferrite toroid). R1 is a 10K-ohm linear-taper carbon composition control, panel-mounted. T1 has a 4:1 impedance ratio (pri has 10 turns of 26 enam. wire on an Amidon FT-50-43 (850 μ u) ferrite toroid. Sec has 20 turns of no. 26 enam. wire. T2 has a 10:1 impedance ratio. Use 30 primary turns of no. 28 enam. wire on an Amidon FT-50-43 toroid. Sec has 10 turns of no. 28 enam. wire.

A BC-band trap may be required between the Beverage antenna and J1 of Fig 5-20 if you live near a strong BC-band AM station, since there is no front-end selectivity at Q1. A simple high-pass filter may be substituted for the trap if several BC stations are operating in your region. See **The ARRL Handbook** for filter tables you may use to design a suitable filter. The cutoff frequency should be 1.7 MHz.

Some amateurs have reported a high noise level with the preamplifier in Fig 5-20. This is an abnormal condition, and is the result of U1 self-oscillation at high gain settings. The usual cause of the malady is improper bypassing at the pins of U1. The bypass capacitors and resistors need to be located as close to the U1 body as practicable, and their leads must be short. If you use an IC socket be sure to install one that is the low-profile type (thin socket with short pins) because the effective length of the U1 pins must be short in order to prevent self-oscillation when so much gain is available.

SIMPLE SUPERHET RECEIVERS

I will not describe elaborate, high-performance superheterodyne receivers in this notebook. After all, this text is aimed at experimenters! Practical examples of high-performance receivers are published in **Solid State Design for the Radio Amateur**, should you wish to become involved with a more comprehensive receiver project.

The principle advantage of a superheterodyne receiver, by comparison with a DC or regenerative receiver, is that the superhet provides what is called "single-signal" reception. There is a signal response at only the upper or lower side of the intermediate frequency. Thus, we can select upper or lower sideband response by changing the receiver BFO frequency, respective to the center of the IF pass-band response. The higher the order of IF selectivity (filtering) the greater the rejection of the unwanted sideband. This feature helps to minimize the effects of QRM that may be present near the desired frequency during CW or SSB reception.

A superhet receiver can be very basic, while still delivering good performance. Such features as an RF gain control, S meter, AGC, digital frequency display, synthesized local oscillator, noise blower or notch filter are not required for most amateur operation. As experimenters and equipment users we can save money and enjoy good performance without the added convenience of the frills provided by commercial equipment manufacturers. If fact, it is not necessary to have an IF-amplifier system in a superhet receiver, provided the overall receiver gain is 75 dB or greater. Fig 5-21 illustrates how simple a superhet can be.

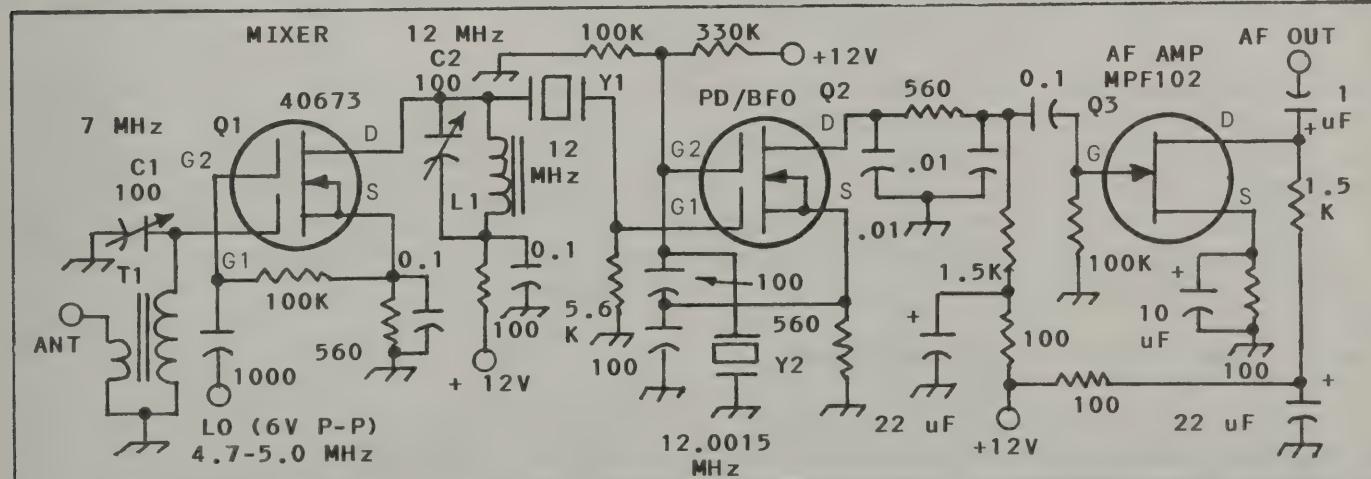


Fig 5-21 -- Circuit example of a minimum-component superhet receiver with no IF amplifiers. Q2 serves as a BFO and product detector. See text for further discussion of this type of circuit.

Dual-gate MOSFETs serve as the mixer and combination product detector and BFO. Q1 and Q2 each have a conversion gain of roughly 15 dB. Y1 functions as a simple IF filter. You may use an IF of your choice by changing the LO frequency accordingly. There are many low-cost computer crystals available for use at Y1 and Y2. The BFO frequency that is determined by Y2 needs to be offset by 1.3 to 1.5 kHz, respective to the Y1 frequency. The BFO must be 1.5 kHz above or below the IF center frequency, depending upon which sideband you wish to copy. This is not an important consideration for CW reception, although most receivers operate in the USB mode during CW reception. Q1 and Q2 may be any dual-gate MOSFET, such as the RCA 40673. The 3N211, 212 and 213 FETs are suitable substitutes for a 40673.

A low-noise first-audio amplifier should follow the product detector, since the audio noise could otherwise override the weaker signals from Q2 of Fig 5-21. A JFET (Q3) serves this purpose. Output from Q3 should be routed to an audio-gain control, and then to an audio amplifier such as an LM386N (see page 61, Fig 3-10).

The tunable local oscillator that provides injection voltage for Q1 of Fig 5-21 must have sufficient output power to ensure a gate no. 2 P-P voltage of 5 to 6 for optimum mixer conversion gain. T1 is tuned to the signal frequency by means of C1. An inductance of 6.5 uH is appropriate for the T1 secondary winding when operating the circuit at 7 MHz (37 turns of no. 28 enam. wire on an Amidon T50-6 toroid core). A 4-turn primary winding may be used for T1. L1 has a 2.2-uH inductance (21 turns of no. 26 enam. wire on a T50-6 toroid).

Although the Fig 5-21 circuit is suggested as a starting point for experimenters, many low-cost refinements are possible. For example, better performance and fewer discrete active devices may be had by using an NE602 doubly balanced mixer in place of Q1. This chip has on-board devices for the local oscillator, which would thereby eliminate the need for an external LO system. A second NE602 can be substituted for the MOSFET at Q2. The internal LO devices would then be used for the crystal-controlled BFO. The combined conversion gain of the two ICs would be approximately 40 dB, which means that only 40 or 50 dB of audio gain would be needed to ensure ample overall receiver gain. Two crystals (1.5 kHz frequency separation) could be used in a half-lattice arrangement in place of Y1 of Fig 5-21. This would improve the IF selectivity.

With the bare-bones approach for a superhet receiver you will discover that front-end mechanical noise (microphonics) are apt to be observed, as is the case when working with DC receivers. Therefore, use care when mounting the mechanical components. Tighten them well and try to use a double bearing variable capacitor for the local oscillator. In any event, you will find little difference in the sound of a received signal with a bare-bones superhet when comparing it against a superhet receiver that has an IF-amplifier system.

Simple Superhet for 20 and 75 Meters

Let's continue following the philosophy of the experimenter. Fig 5-22 illustrates a practical bare-bones superheterodyne receiver that can be used on 20, 75 and 80 meters. The local oscillator operates from 5.0 to 5.4 MHz. This frequency range results in USB reception on 20 meters and LSB reception on 75 meters without changing the BFO frequency. It's a very old scheme that dates back to the early days of SSB.

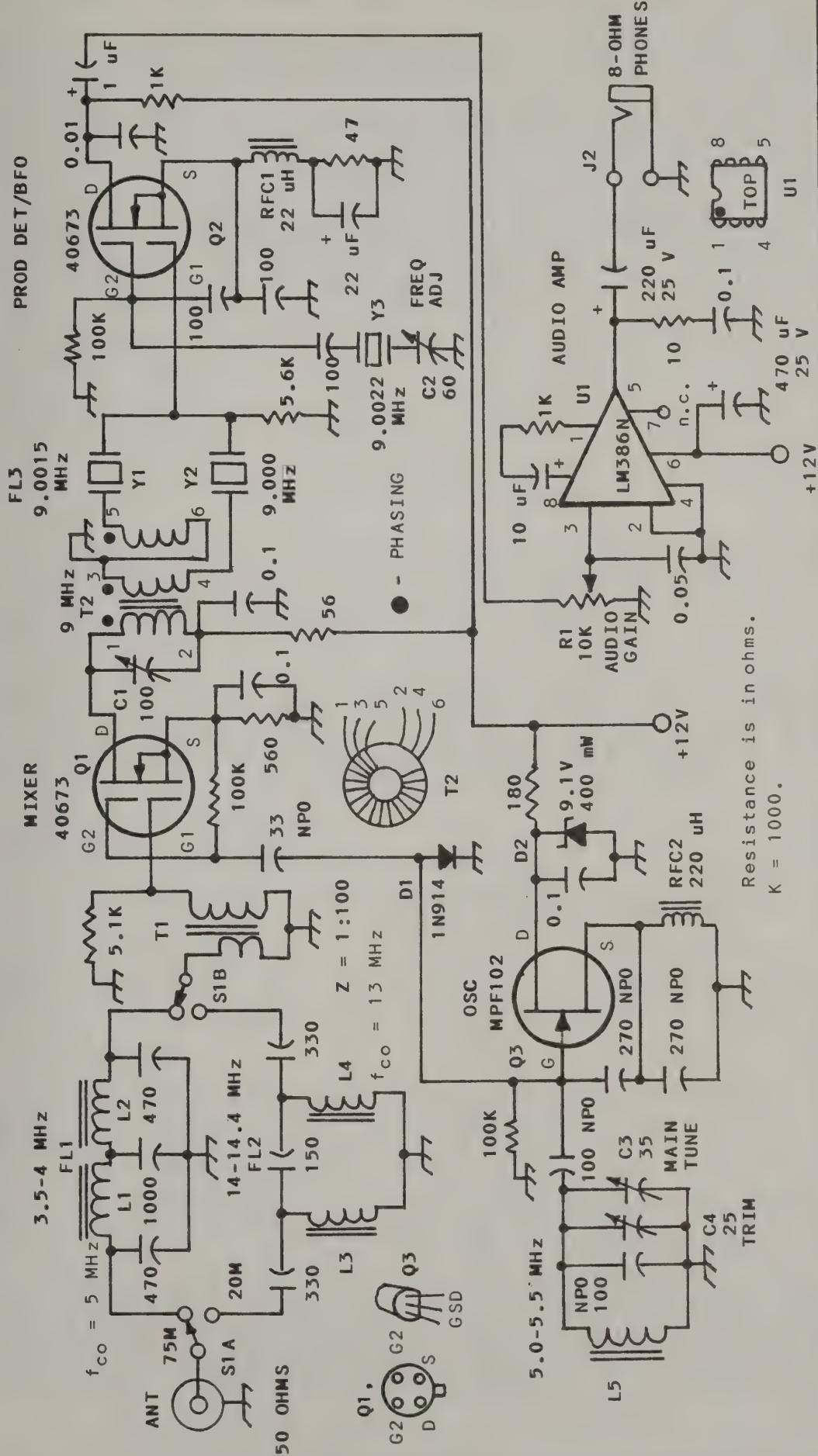


Fig 5-22 -- Schematic diagram of a minimum-component superhet receiver for 20 and 75 meters. Capacitance is in μ F except decimal-value units, which are in μ F. Polarized capacitors are tantalum or electrolytic. Resistors are 1/4-W carbon film or carbon composition. C1 is a mica trimmer. C2 is a plastic or ceramic trimmer. C4 is an NPO ceramic trimmer. FL1 and FL2 capacitors are polystyrene or silver mica. NPO = zero temperature coefficient. R1 is an audio taper carbon composition control. See text for coil data.

L1 and L2 in Fig 5-21 are 2.0-uH inductors. Use 20 turns of no. 24 enamel wire on an Amidon T50-2 toroid. L3 and L4 are 0.47-uH inductors. They require 11 turns of no. 24 enamel wire on a T50-6 toroid core. VFO coil L5 is 4.8 uH. Wind 32 turns of no. 22 enamel wire on a T68-6 toroid core, then apply at least two coatings of Polystyrene Q Dope or an equivalent high-Q coil cement. Allow 24 hours between coil-dope applications.

T1 is a broadband transformer that matches the 50-ohm filters (FL1 and FL2) to 5100 ohms (via Q1 gate resistor). The T1 secondary has 20 turns of no. 24 enamel wire on an Amidon FT-50-43 ferrite toroid. Use 2 turns of no. 24 enamel wire for the primary winding. T2 is a trifilar-wound inductor that has an inductance of 4 uH. Use 28 trifilar turns of no. 28 enamel wire on a T50-2 toroid core. The wires may be first twisted together with a hand drill to provide roughly 8 twists per inch. Be sure to connect the T2 leads according to the polarity indicated.

Main-tuning capacitor C3 is an air variable type that turns freely. Use a vernier drive to provide ease of tuning.

Circuit Description

The Fig 5-22 circuit has no frills. It is designed for maximum performance with minimum parts. A low-pass filter (FL1) is switched in by S1 during 75- and 80-meter operation. The low-pass configuration keeps 20-meter signals from being received along with the 75-meter ones. It is designed to start attenuating at 5 MHz. FL2, on the other hand, is a high-pass filter that allows signal energy above 13 MHz to reach Q1. It prevents 75-meter signals from being heard while listening on 20 meters. Very strong local signals may still leak through from the unwanted band. If this becomes a problem, simply add a 20-meter parallel trap between S1A and FL1. In a like manner, add a 75-meter trap between S1A and FL2, should a nearby amateur disrupt your 20-meter reception.

Oscillator injection voltage for Q1 is taken directly from the gate of Q3. This circuit point has a nearly pure sine wave (minimum harmonic current), which ensures minimum spurious response from the mixer. D1 serves as a bias stabilization diode for Q3, which in turn minimizes oscillator drift.

You will observe that there are no IF amplifiers in the receiver. An overall receiver gain of approximately 70 dB is available, and this yields more than ample audio output into 8-ohm headphones. Insertion loss through the IF filter, FL3, is on the order of 3 dB.

Product detector Q2 serves also as the BFO. This technique proved successful in the W1FB 75-meter receiver described in **QST** for May 1989, page 25 ("A Four-Stage 75-Meter SSB Superhet"). C2 adjusts the crystal frequency to the desired point on the IF response curve. It is adjusted for minimum response of the unwanted sideband, consistent with voice fidelity and minimum loss of receiver gain. A 1N914 diode may be added from gate no. 2 to ground at Q2. This will help reduce harmonic currents in the BFO portion of the stage.

If you desire to use a speaker with this receiver it will be necessary to use an audio preamplifier between Q2 and U1. Suitable circuits are described in the early chapters of this book. Audio shaping is accomplished by way of the 0.01-uF capacitor at the drain of Q2, and by virtue of the 0.05-uF capacitor at pin

3 of U1. These capacitors, plus the resistance in R1, do double duty by keeping BFO energy at the Q2 drain from reaching U1.

Local oscillator Q3 has regulated voltage applied to its drain. This aids VFO stability. There is more than enough RF voltage at the Q3 gate to provide the required 5-6 volts of P-P injection voltage for gate no. 2 of Q1. The 33-pF series capacitor from Q1 to Q3 may be changed to obtain the desired injection voltage, should your circuit have too much or too little VFO voltage at gate no. 2 of Q1. Measurements of this voltage are done with a scope and a X10 probe. If you don't have a scope and probe you may use a VTVM and an RF probe. You will need 2.12 volts rms at gate no. 2 of Q1 with this measurement technique.

Construction Tips

It is important that you keep all RF signal leads as short and direct as possible in order to assure high gain and to avoid self-oscillations. The universal breadboard described earlier in this book may be used as a foundation for the receiver. Certainly, it should not be a major challenge to lay out your own PC board for this project, owing to the small number of parts used.

Improved performance may be had by following the suggestion for the circuit of Fig 5-21, where I mentioned the possible use of NE602 ICs in place of Q1 and Q2 in that circuit. The overall gain would be greater, and the mixer performance (dynamic range) would be vastly improved.

You should be aware that when using the frequency scheme specified for the Fig 5-22 circuit, the tuning will be "backwards" on one of the bands. To be concise, the low end of 80 meters will be at one end of the main-tuning dial, but the low end of 20 meters will be at the opposite end of the readout dial. It is for this reason that separate calibration scales are necessary.

Receiving the Other HF Bands

The Fig 5-22 circuit may be used for reception on 10, 12, 15 and 17 meters by using a converter ahead of the receiver. The tunable IF (main receiver) becomes 3.5 to 4.0 MHz. The converter can be built as a separate module that fits into the cabinet of the receiver. You may prefer to have a converter that covers one additional HF band, or you may elect to band switch your converter to provide multiband coverage. Alternatively, you may choose to build a separate converter for each additional band of interest. The converters may be selected by means of a band switch if this is done. Switching is less difficult under the latter condition, because all of the switching is done at low impedance (50 ohms). A three pole, multiposition wafer switch permits switching the converter input and output lines, along with the +12-V supply lines.

Fig 5-23 shows the circuit for a minimum-parts HF converter. Only two transistors are used in this circuit -- a mixer and an overtone crystal oscillator. L2 is the high-Q input tuned circuit. Only two values of inductance are required to provide coverage of the 10, 12, 15 and 17 meter bands. In other words, two bands are accommodated by one inductance value. A low-pass network matches the mixer output impedance (established by the 1.2K-ohm resistor) to the 50-ohm input of the receiver. The low-pass network helps to prevent Q2 oscillator energy from being passed to the input of the main receiver.

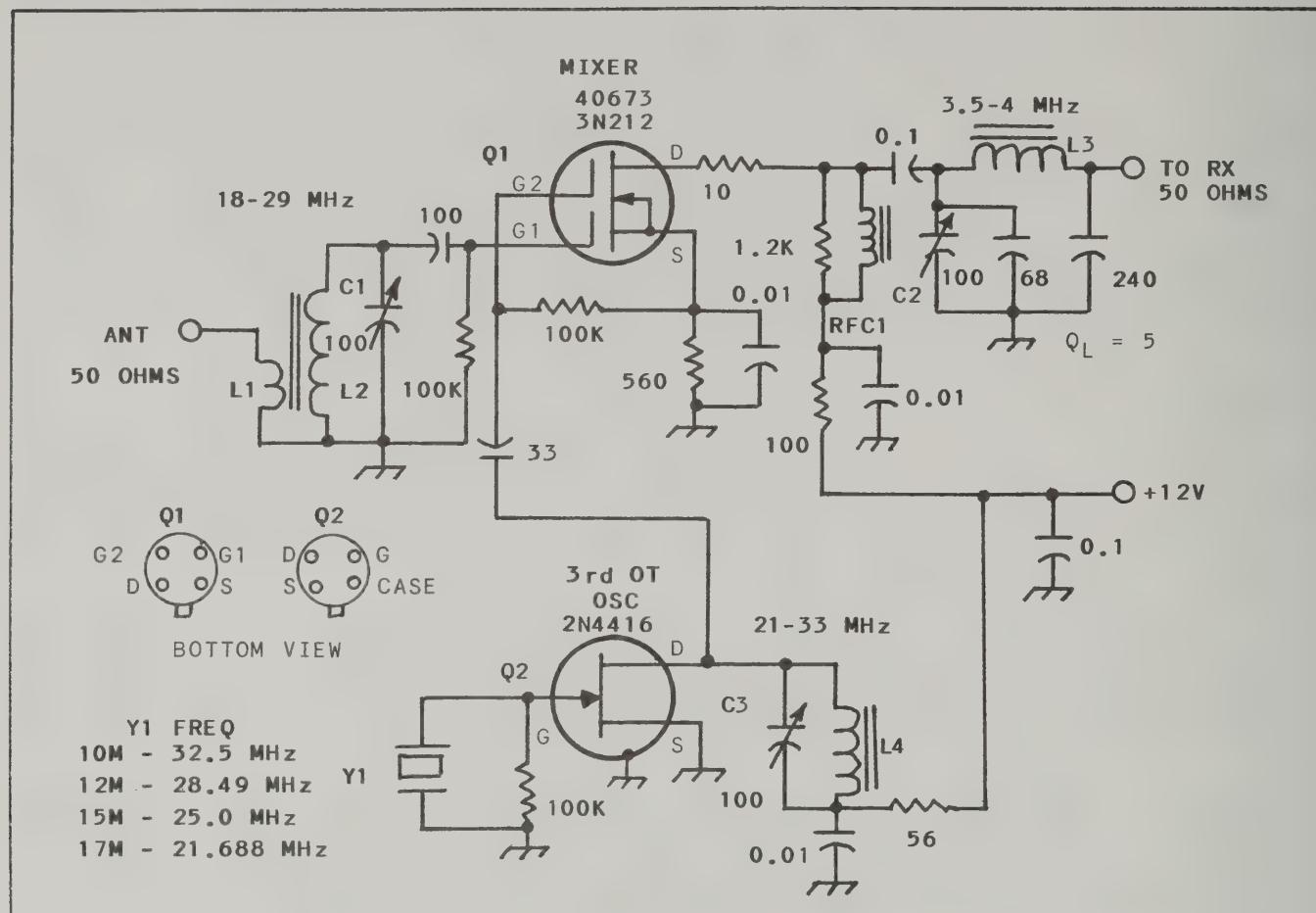


Fig 5-23 -- Schematic diagram of an HF-band converter that may be used with the receiver in Fig 5-22 to cover the 10, 12, 15 and 17 meter bands. Fixed-value capacitors are disc ceramic. Resistors are 1/4-W carbon film or carbon composition. C1, C2 and C3 are 100-pF ceramic or mica trimmers. L2 is 0.5 μ H for 10 and 12 meters (10 turns of no. 24 enam. wire on an Amidon T50-6 toroid). L2 is 0.95 μ H for 15 and 17 meters (14 turns of no. 24 enam. wire on a T50-6 toroid). L1 has two turns of no. 24 enam. wire. L3 is a 10.2- μ H inductor (45 turns of no. 30 enam. wire on a T50-2 toroid). L4 has 0.5 μ H of inductance for 10 and 12 meters (13 turns of no. 24 enam. wire on a T37-6 toroid). L4 is a 0.7- μ H inductor for 15 and 17 meters (15 turns of no. 24 enam. wire on a T37-6 toroid). RFC1 is a miniature 1-mH RF choke. Y1 is a third-overtone crystal, 20-pF load capacitance.

C1 is adjusted for peak signal response at the center of the band of interest. C2 is adjusted for maximum signal at 3.75 MHz. C3 is set for reliable starting of the oscillator. Generally, the tuned circuit is set for a frequency that is slightly higher than the marked crystal frequency.

This converter is by no means a monument to high gain and low noise figure. The noise figure may be improved by adding a low-noise RF amplifier ahead of the mixer. However, this circuit delivers adequate performance for all but the weakest of signals.

The response of the L3 network may be broadened by placing a 3.3K-ohm resistor in parallel with L3. This will reduce the converter gain slightly.

Chapter Summary

We have covered the fundamentals of receivers and viewed some simple but practical circuits for receiving. Many of the circuits I have presented can be massaged to greatly improve their features and performance. Active mixers have been highlighted to provide conversion gain as a contribution to the overall receiver gain. Doubly balanced diode mixers are better performers, but they require higher levels of LO injection (+7 dBm or better), which creates the need for an RF amplifier after the oscillator. Also, the NF of diode mixers is high, which dictates the need for an RF preamplifier ahead of the mixer. Owing to the conversion loss of diode mixers, it is necessary to employ IF amplifiers after the mixer. This is definitely the correct route to travel when we build high-performance receivers. But, for the purpose of learning and building simple equipment, we can enjoy many hours of acceptable performance with the circuits presented here. You will find a more elegant approach to receiver design in **Solid State Design for the Radio Amateur**.

The Universal Breadboard described earlier in this book is suitable as a foundation for the circuits in this chapter. These boards may be purchased from N9ATW for a nominal fee.

TRANSMITTER DESIGN & PRACTICE

CHAPTER 6

We should not treat solid-state transmitter design in a casual fashion. Although the circuits we use may seem simple compared to vacuum tube ones, we must play an entirely different game if we are to expect good performance. Some amateurs try to use vacuum-tube principles when they design a solid-state transmitter. In particular, they attempt to use high-impedance tuned circuits in circuit locations where the impedance is typically very low. Whereas a 5-watt vacuum-tube RF amplifier may have a plate impedance of several thousands of ohms, a 5-watt transistor RF amplifier may have a collector impedance as low as 10 ohms. The severe mismatch can destroy a transistor quickly, and even if the transistor could endure so hostile an environment, the available output power would be reduced drastically because of the mismatch.

Inexperienced designers generally short-change themselves by not including adequate harmonic filtering for their solid-state transmitters. The requirements are more stringent for semiconductor devices than they are when working with tubes. Harmonic energy is generated in tubes by way of envelope distortion. Transistors not only produce harmonic currents by this means, but also from the inherent varactor action within the transistor junction. This nonlinear change in junction capacitance is well known for harmonic generation. It is not unusual to find harmonic currents at the unfiltered output of a solid-state transmitter a mere 10-15 dB below the desired fundamental energy. Should we worry about this? Definitely! TVI and other forms of RFI are a potential threat, even when operating a QRP rig. High levels of harmonic energy can ruin the efficiency of a transistor amplifier when the driving energy is not reasonably pure. The net effect is that the driven stage draws more current than it would when excited by a clean waveform, but the output power (fundamental) does not increase in proportion. A transistor amplifier that is driven by a dirty signal may run very hot, whereas if excited by a clean signal it will operate in cool manner. Harmonic filtering is simple and inexpensive. It is also required by FCC regulations.

Gain distribution is also an important factor in transmitter design, just as it is in a well-designed receiver. I have seen a number of home-made transistorized transmitters that were literally gasping for air because some of the stages were so severely overdriven. Transistor manufacturers rate their RF devices for "saturated output power." Saturation occurs when no additional output power occurs with an increase in excitation power. As the driving power is increased beyond saturation, the collector current rises without an attendant change in RF output power. Bingo! The amplifier efficiency goes sour. Furthermore, harmonic energy increases beyond the saturation state, and the safety of the transistor becomes suspect under these adverse conditions. If I may offer a rule of thumb at this juncture, I suggest that you check each transmitter stage for saturation. Use only enough driving power to reach the saturation point, then reduce the drive slightly to just below the saturation condition. This may be done with an RF probe and VTVM, or with a scope. Be sure that each stage checked is terminated by its proper load.

Insufficient driving power is another common difficulty. With vacuum tubes we always knew how many grid no. 1 mA were required to drive a particular tube to its full rated output power. The situation is somewhat different when we design with transistors. Devices that are designed for RF power service are rated for a specific driving power in watts or mW. This data is provided on the data sheets that are provided by the manufacturer. For example, a particular power transistor may be capable of delivering 10 watts of RF power (P_o). The driving power required might be 1 watt (P_{in}). This equates to a device power gain of 10 dB. If we know the specified gain of a given power transistor, but do not have the driving-power data, we can approximate the required driving power by knowing the gain and output power of the device. It is important to have slightly more driving power available than is specified. This will compensate for losses in the coupling network between the driver stage and the power amplifier. If the transistor I plan to use calls for, say, 500 mW of P_{in} , I design the driver for at least 750 mW of output power.

We amateurs frequently work with transistors that are not earmarked for RF power service. Many hi-fi audio power transistors have high f_T ratings, and they lend themselves well to RF service. It is necessary at times to "fly by the seat of one's trousers," as the saying goes, to determine how much RF input power is needed to make the audio device percolate well at RF. This can usually be determined by experimenting with the driving power. Find the point where saturation occurs, then reduce the driving power slightly below that value. Such audio transistors as the MPSU02 and MPSU05 are examples of low-cost devices that perform nicely as RF power transistors up to 30 MHz.

Once again I want to stress the importance of correct impedance matching between the stages of a transmitter. A significant mismatch condition will reduce the available output power from the driver stage or earlier stages. Even though there is ample power otherwise available, too little excitation power may reach the stage being driven because of the mismatch. Adjust the turns ratio of the tuned circuit or broadband transformer between stages for optimum excitation if you do not know the load impedance of the stage being driven. Don't be afraid to experiment. After all, that's what ham radio is all about!

Instability Problems

Self-oscillation is a common problem in home made transistorized transmitters. Generally speaking, stability is easier to achieve with transistors than it is with tubes. This is because the input and output impedances of transistors are substantially lower than they are with vacuum tubes. Nonetheless, transistors will "take off" if certain precautions are not taken when designing a circuit. A contributing factor to instability is the phenomenon associated with an increase in device gain as the operating frequency in MHz is lowered. This encourages self-oscillation at LF, VLF and even at audio frequencies. Therefore, we must ensure that bypassing of the voltage (V_{cc}) supply lines is adequate also for low frequencies. This is why you generally find a 22-uF or similar value capacitor in parallel with, say, a 0.001- and a 0.1-uF bypass capacitor at key points in an RF power circuit. The use of shunt feedback (collector to base) also serves to prevent self-oscillation in class A and AB types of RF amplifiers.

Circuit-board layout, of course, plays a role in amplifier stability also. I prefer what I call "in-line" layout. The objective is to isolate, as much as possible, the input of each stage from the output port, and to keep each stage as far away from the previous and following stages, physically, as necessary.

In the case of grounded- or common-emitter amplifiers it is important to keep the emitter ground lead short and direct. The PC-board foil to which the emitter connects should be as wide and direct (to the main ground bus) as practicable. This minimizes stray inductance, which can cause instability and a loss of amplifier gain (degeneration). This consideration is not so important when using push-pull power transistors because the RF current is flowing from emitter to emitter, and the PC board emitter foil simply provides a dc ground. Push-pull transistors should, however, be mounted as close to one another as possible in order to keep the emitter leads short.

Self-oscillation generally occurs at low drive levels. As the collector current increases at higher driving power, a self-oscillation may cease. The usual cure for this malady is to reduce the Q of the amplifier input circuit. Resistive swamping is often used for this purpose (base to RF ground). The lower the resistor value the greater the driving power needed to fully excite the amplifier, since driving power is dissipated in the de-Qing resistor. If an RF choke is used as a base return element, try placing an 850 mu ferrite bead on one or both pigtails of the RF choke. This lowers the choke Q to 10 or less. Alternatively, the RF choke may be used in parallel with a low-value resistor. This also lowers the Q of the choke. Fig 6-1 illustrates some of the techniques we have treated here.

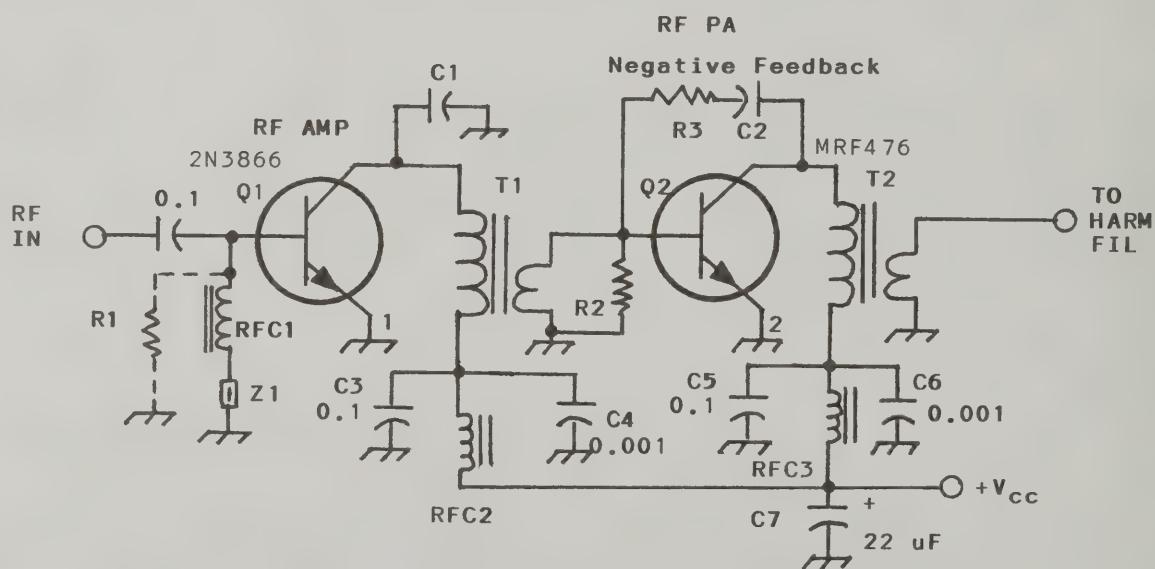


Fig 6-1 -- Example of a broadband class C RF amplifier strip that has been treated for instability. Circuit points 1 and 2 (emitter) require short, direct ground connections. Z1 is an 850 μ i ferrite bead that lowers the Q of RFC1. R1 or a combination of R1 and Z1 may be used for the same purpose. A typical value for R1 is 10 to 100 ohms. R2 is used also to lower the Q of the input transformer, T1. A ferrite bead is sometimes slipped over the T1 secondary return lead to lower the winding Q. C4 and C6 are VHF bypasses. C3 and C5 are effective bypass capacitors for HF and MF. C7 is for bypassing LF, VLF and audio energy. C1 damps VHF self-oscillations and helps remove VHF harmonic currents. R2 is 10 to 47 ohms for HF. See text for further comments about this circuit.

RFC1 in Fig 6-1 should have an inductance value that corresponds with an XL that is four to ten times the characteristic base impedance of Q1 at the lowest operating frequency. For example, if the base impedance is 15 ohms and the operating frequency (lowest) is 3.5 MHz, RFC1 should have no less than 2.72 uH. A 5- or 10-uH choke will work well in this example. RFC2 and RFC3 are part of the Vcc decoupling circuit, which prevents RF energy from migrating along the supply line from Q2 to Q1 or vice versa. Unwanted RF currents on the supply line, if of the proper phase, can cause self-oscillation. RFC2 and RFC3 need not have high inductance because the Vcc line has a very low impedance, assuming the power supply is designed correctly. A 5-uH inductance is adequate. These two chokes must have wire that can accommodate the current drawn by Q1 and Q2 without a resultant voltage drop across them. For the Fig 6-1 circuit you may use 6 turns of no. 22 enamel wire on an Amidon FT-37-43 ferrite toroid.

T1 and T2 in Fig 6-1 are broadband transformers that are wound on type 43 (850 ui) ferrite toroids. The turns ratios are adjusted to provide an impedance match between the transistors and their loads. The secondary windings of these transformers, or whichever winding interfaces with the lowest circuit impedance, need to have an inductance that equals an XL that is four times or greater the load impedance (as with RFC1 of Fig 6-1).

C2 and R3 in Fig 6-1 form a negative feedback network that helps stabilize Q2. The values for these components may be determined experimentally. In a typical case we might have a C2 capacitance of 1000 pF and an R3 resistance of 500 to 1000 ohms. The values should be chosen to provide only enough feedback, collector to base, that ensures stability and flat gain across the amplifier operating range in MHz. Remember that the greater the feedback the lower the amplifier output power. I generally design for one to two dB of feedback power when I work with high gain power transistors (13 dB or greater gain). I use less than one dB of feedback power for transistors that have lower gain. Negative feedback may be applied to any RF amplifier, irrespective of the operating class.

A common cause of amplifier instability can be traced to a mismatched or reactive load. This means that we must ensure that the transistor matches the harmonic filter, and that the filter is terminated by the correct resistive load impedance. The same is true of low-level RF stages. They need to be matched to one another. There are situations, however, where a designer may introduce an intentional mismatch between amplifier stages in order to establish suitable gain distribution. This practice provides an invitation to instability. It is far better to alter the stage gain by changing the transistor bias or through the use of negative feedback.

Thermal Considerations

Heat is the enemy of most electronics components, and transistors are no exception. As the transistor junction temperature rises above the manufacturer's maximum allowances the device gain can increase. With this elevated gain comes higher collector current, and eventually self-destruction (thermal runaway). It is important that we provide a heat sink of sufficient area to keep the transistor temperature at or below the maximum rating. No transistor should be allowed to become so warm that one can't hold his finger on the case for a period of time. If the transistor body is too hot to handle, it's too hot to operate safely.

The transistor has to be well bonded thermally to the heat sink in order to ensure the effective transfer of heat. A thin coating of transistor heat-sink compound

(silicone grease and zinc oxide) should be placed on the surface of the transistor that mates with the heat sink. A snug fit (but not excessively tight) between the transistor and the heat sink is also essential. Some commercial equipment uses forced-air cooling to keep the heat sink and the transistors at safe operating temperatures. Needless to say, the heat sink won't be damaged, but it will do its job better if it is kept cool.

Linear solid-state amplifiers require larger heat sinks than do class C amplifiers. This is because a class-C amplifier draws only microamperes of current when no signal is present. Linear amplifiers, on the other hand, draw idling or standing collector current during no-signal conditions. This current causes heat, and the heat sink does not cool down as effectively as it does in class-C operation. Also, the operating duty cycle is often greater during SSB operation than it is for CW service. This provides less opportunity for cooling during the overall operation of the equipment. RTTY operation is especially stressful respective to duty cycle and transistor heating. Likewise with AM operation. I recommend that you reduce the dc input power to your solid-state amplifier by 50% when using the AM or RTTY modes. If not, your transistors may have short lives! In fact, it is not uncommon to reduce the power by 75% for these modes. It depends upon the power-output capability of your transistors and the effectiveness of the heat-sinking hardware.

The effectiveness of a heat sink depends largely on the case style of the transistor used with it. A TO-3 type of transistor has greater area for mating to the heat sink than does, for example, a TO-220 device. Also, the material from which the heat sink is made plays an important role in heat reduction. Aluminum is the preferred metal for heat sinks, although some (such as the TO-5 press-on sinks) are made from brass. It is believed generally that a flat black anodized or painted heat-sink surface improves the thermal quality of the sink, compared to bare aluminum hardware. In any event, the heat-sink surface must be as smooth and flat as possible at the point of transistor contact. If not, heat transfer will not be effective.

The RF PC Board

Electrical stability may be enhanced by using a double-sided PC board for RF amplifiers. A double-sided board is one that has copper on both sides. One of the surfaces is etched in accordance with the desired PC or etched pattern. The remaining side is left intact to serve as a ground plane for the etched side of the board. This practice reduces circulating currents that can otherwise create what are known as "ground loops" or RF hot spots along the ground foils of the PC board. The ground foils on the etched side of the board are made common to the ground plane. The etched conductors thus form small capacitors with the ground plane, with the PC-board insulating material serving as the capacitor dielectric. These capacitors do not hinder the circuit performance from audio through the HF spectrum, since they are small in value (5 to 30 pF typically). The normally low-impedance circuits on the PC board are not affected significantly by these capacitive reactances. A benefit of this parasitic capacitance is the bypassing of VHF and UHF currents that might otherwise encourage VHF self-oscillation.

I do not recommend that you use double-sided PC boards for VFO or other stability-dependent circuits. The parasitic capacitors with their phenolic or fiber-glass dielectric are not temperature stable. They can cause severe oscillator frequency drift. Furthermore, flexing or vibration of the PC board can alter the value of these parasitic capacitors and affect frequency stability.

Dealing with Power MOSFETs

Owing to the fragile nature of the insulation between the gate and drain-source junction of MOSFETs, it is necessary to treat these devices with special care. Static charges can perforate the gate insulation easily if the transistor is not held in conductive foam plastic until it is installed in the circuit. This is not necessary for power MOSFETs that have internal gate-source protective Zener diodes. It is wise to ground the soldering iron tip while soldering a MOSFET to its PC board. The transistor should be the last component you place on the board. Avoid using excessive heat when soldering the gate pin of the device.

Power MOSFETs are sensitive to excessive gate voltage and current in an operating circuit. The peak-to-peak voltage at the transistor gate should not be allowed to exceed 30 under any circumstances, lest the gate insulation become damaged. Signal-voltage transients or spikes, no matter how short the duration, can destroy the transistor instantly if they exceed the 30-V value. This causes a gate-source short circuit. For this reason it is important to ensure that the low-level stages that precede the MOSFET amplifier are free of spurious oscillations and switching transients. The same is true of the forward-bias voltage that is applied to the FET gate. External back-to-back Zener diodes may be bridged from gate to ground as a protective measure. I generally use a pair of 15-V, 400-mW Zeners for this purpose. The additional shunt capacitance from the diodes does not seem to create difficulties at MF and HF. The situation is more critical at VHF.

Excessive voltage, sustained or momentary, from drain to source can also destroy a MOSFET quickly. Some of the modern power switching FETs contain a built-in drain-source Zener diode. One may be used externally if the transistor does not have one built in. It should be rated for twice the drain supply voltage to allow for the sine-wave excursion in an RF or audio circuit. I use a 50- or 55-V Zener diode for a +24-V supply.

Power MOSFETs are especially susceptible to VHF self-oscillation. This is because RF power MOSFETs in particular require an internal geometry that makes them work well up to approximately 175 MHz. This malady can be prevented by de-Qing the gate circuit. A good practice is to place a 10- or 15-ohm, 1/4-W carbon film or carbon composition resistor in series with the energy supplied to the gate. The resistor should be located at the gate terminal. It is sometimes helpful to use a ferrite bead over the lead of a 10-ohm resistor for this purpose. Another technique for discouraging self-oscillation is to keep the gate at low impedance. The characteristic gate impedance of a MOSFET is one megohm or greater. The effective gate load is established by means of the gate resistor used. I try to keep the impedance at 500 ohms or lower. This requires additional power from the driver stage in order to develop the gate-voltage swing that is needed to turn on the transistor. Power MOSFETs are enhancement-mode devices. This means that a specific gate voltage is required to turn them on. This is not the case with depletion-mode transistors such as the JFET (MPF102 for example).

Although the foregoing commentary may seem like an indictment of power MOSFETs, it's not the case. Unlike bipolar transistors, the MOSFET internal capacitances do not change with variations in operating voltage, frequency and drive power. This makes it an easy matter to design a feedback network that works well across a wide range of frequency. Input and output matching transformers or networks are also predictable in terms of performance versus frequency. Power MOSFETs are immune to thermal runaway, unlike bipolar transistors.

An uncomplicated gate-bias system is acceptable for power FETs because the gate draws only μ A of current. This makes regulation of the bias voltage unnecessary. An ordinary resistive divider that is connected to the drain supply (V_{dd}) can be used successfully. A forward bias of approximately +1 V is required to turn on a power MOSFET. A bias as great as +3 V is commonly applied to the FET during linear service. The desired resting drain current, which varies with the device used and the class of service, is established by means of the forward-bias value. A resting drain current of 100 to 200 mA is typical for class AB operation.

Another plus feature of power FETs is their efficiency. Whereas a class AB bipolar transistor amplifier in broadband service has an efficiency of roughly 50%, a power FET in the same operating class may yield an efficiency as great as 65 or 70 percent. I have built class C FET amplifiers that had efficiencies as great as 80%. Ed Oxner, KB6QJ, who is an engineer at Siliconix, Inc. once told me that his colleagues at Siliconix had reported 90% efficiency with some MOSFET amplifiers they had developed in the lab. Helge Granberg, K7ES, who is an RF applications engineer at Motorola in Phoenix, reported to me some years ago that the IMD (intermodulation distortion) products, especially the high-order ones, were substantially lower in a FET amplifier than is typical with a bipolar-transistor amplifier.

Many audio and dc switching power FETs are suitable for RF use in the HF spectrum. The IRF511 is an example of a low-cost FET that can deliver up to 15 watts of RF output power when operated from a +24 volt power supply. However, the performance of some dc and audio types of power FETs falls off markedly above 10 MHz. Although they can be driven easily to full drain current, the output power sags miserably, and this spoils the efficiency.

PRACTICAL TRANSMITTERS

A small solid-state CW transmitter can be built more quickly than a vacuum-tube equivalent model. There is no need to drill holes in a chassis, mount large hardware items and purchase costly tubes (and air variable capacitors!). A circuit can be tacked together on an unetched scrap of PC board material. The field for experimentation is wide open for those of you who have the curiosity and zest for this rewarding pastime. I can't stress enough the thrill of communicating with another ham while using something that you built from scratch.

If you have not attempted to construct a solid-state transmitter, I recommend that you commence with a one-transistor crystal-controlled oscillator. This will help to enhance your self-assurance, and you will gain knowledge in the process. An amplifier stage may be added later to boost the power of your transmitter. Figs 6-2 through 6-4 contain circuits for a progressive CW transmitter of this type. Build one stage at a time, use it on the air, then move along to the next step. The universal breadboard described earlier in this booklet may be used for your circuits. Build the first stage (crystal oscillator) in one corner of the board. This will provide ample space for the two amplifiers that follow.

I have specified the 40-meter band for this project, since 7 MHz is generally open both day and night. Furthermore, there is plenty of CW activity on 40 meters. You should have no difficulty getting someone to answer your CQs if you use an effective antenna, such as a dipole, end-fed wire or vertical that is matched to the feed line and transmitter.

Fig 6-2 contains the circuit for stage no. 1 of the transmitter. The RF output from this circuit is on the order of 1/4 watt, which will net you many QSOs.

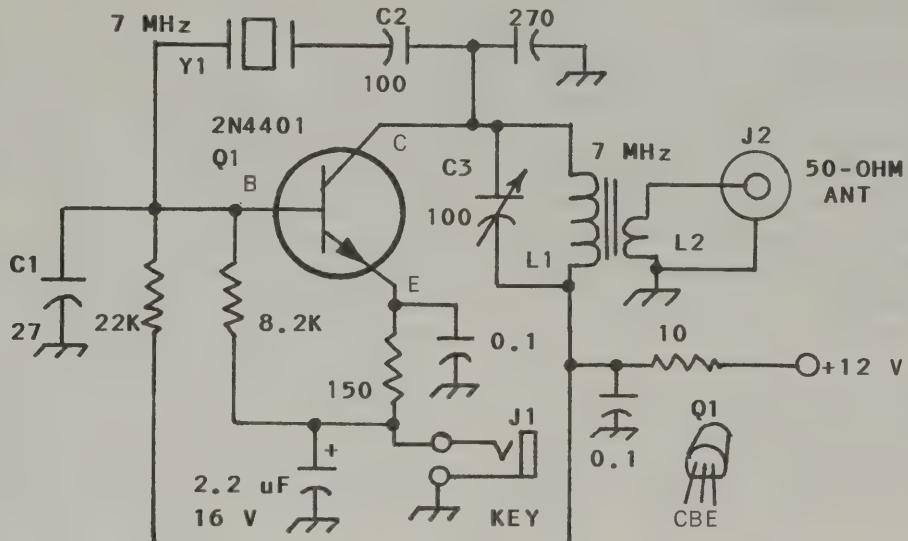


Fig 6-2 -- Schematic diagram of a 150-mW CW transmitter. Fixed-value capacitors are disc ceramic, 50-V or greater. Resistors are 1/4- or 1/2-W carbon film or carbon composition. C1 and C2 are feedback capacitors (see text). C3 is a 10-100 pF ceramic or mica compression trimmer. A two-circuit phone jack is used for J1 and an RCA phono connector or SO-239 coax connector may be used for J2. L1 is a 2.3-uH toroidal inductor. Use 24 turns of no. 26 enam. wire on an Amidon T37-6 toroid core. L2 has three turns of no. 26 enam. wire wound over the +12-V end of the L1 winding. Q1 is a 2N4400 or 2N4401 transistor. A 2N2222A may be substituted, but will deliver less output power. Y1 is a fundamental crystal, 30 pF load capacitance.

You should have no difficulty getting the above circuit to work properly. If Y1 does not oscillate on the first try, adjust C3 until oscillation occurs. If this fails to remedy the problem, and assuming the crystal is okay, increase the value of feedback capacitor C2 slightly. Use no more capacitance at C2 than is necessary to ensure reliable oscillation. Too much feedback can cause spurious output energy and a chirpy CW note. C3 needs to be adjusted for a chirp-free CW note when keying the transmitter. Initial testing should be carried out while using a 1-W, 47- or 56-ohm resistor as a dummy antenna.

The Fig 6-2 circuit may be used on 20 and 30 meters by reducing the inductance of L1 accordingly. The 270-pF capacitor at the junction of C2 and L1 will need to be reduced to 100 pF for 20-meter operation. Y1 should be chosen for the desired 20-, 30- or 40-meter operating frequency.

You may use a 2N3866 or equivalent transistor at Q1 in Fig 6-2 if you desire slightly more output power. A 2N3866 should yield approximately 1/4 watt of power output. A small increase in power is possible if you reduce the value of the Q1 emitter resistor. Do not use a value below 68 ohms, lest the transistor draw excessive current and become defective. You may wish also to experiment with the turns ratio of L1 and L2 toward obtaining maximum oscillator output power, consistent with a non-chirpy CW note. This may be done by changing the number of L2 turns. More turns results in tighter coupling to the antenna. An RF probe or scope, plus a dummy antenna, may be used for this test.

Adding a Driver/Amplifier for Increased Output Power

Phase 2 of our beginner's transmitter calls for the addition of an RF power stage that you can use to extend your communications range. This part of the transmitter is arranged for class-A linear service. Although it is not necessary to use a linear amplifier for CW or FM amplification, there are some advantages: (1) a linear amplifier produces a lower level of harmonic currents; (2) it is easier to drive when a low-power stage is used to excite it; (3) the keyed waveform of the overall transmitter is less clicky than when using a class-C amplifier after the keyed stage.

The second stage of our transmitter is described in Fig 6-3. Output power from Q2 is approximately 1 watt. This level of power will enable you to work DX when band conditions are good.

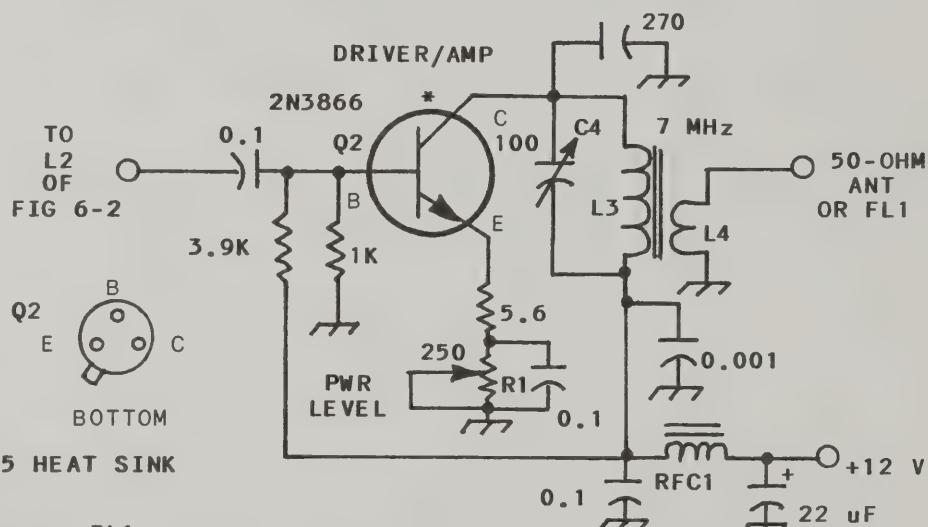


Fig 6-3 -- Schematic diagram of the linear driver/amplifier for the progressive CW transmitter. Fixed-value capacitors are disc ceramic except where otherwise specified. Resistors except for R1 are 1/4-W carbon film. C4 is a ceramic or mica compression trimmer. FL1 is a half-wave harmonic filter with a low-pass response. Cutoff freq. is 7.8 MHz. C5, C6 and C7 are silver mica or polystyrene. L5 and L6 of FL1 are 1.2 μ H. Use 15 turns of no 24 enam. wire on an Amidon T50-2 toroid (see text). Q1 may also be a 2N3553 or 2SC799 transistor. R1 is a 0.5-W carbon composition control, linear taper. RFC1 is 6 turns of no. 24 enam. wire on an Amidon FT-37-43 ferrite toroid. L3 of T1 is a 2.3- μ H inductor. Use 24 turns of no 26 enam. wire on an Amidon T50-6 toroid. L4 has 3 turns of no. 26 enam. wire over +12-V end of L3 winding. The polarized 22- μ F capacitor is electrolytic or tantalum.

You will note that a harmonic filter is shown at the lower left in Fig 6-3. Even though the amplifier circuit can be used without the filter, I strongly recommend that you include it when operating into an antenna. This will lessen the chance for TVI and will allow you to adhere to the FCC regulations for spectral purity. The filter can be used also when you use Q2 to drive the third stage of our transmitter. It will assure a clean waveform at the input of the final amplifier stage.

R1 in Fig 6-3 has been included for use as a drive control when you add the last amplifier stage. It will permit varying the transmitter power over a wide range for those special QRP QSOs you may desire. It is used also to prevent driving the third transmitter stage beyond the saturation point. R1 may be increased to 500 ohms if you desire greater power range.

If you should experience instability with Q2, try placing a 3.3K-ohm, 1/2-W resistor in parallel with L3. This will reduce the power output slightly because some of the RF energy will be dissipated in the resistor.

Keep all signal leads as short and direct as practicable. Place the Q2 circuit at least an inch away from the Q1 circuitry to reduce unwanted stray coupling between the circuits. A push-on finned TO-5 heat sink is required on Q2 to keep it from overheating. Use a thin coating of heat-sink compound or clear silicone grease between the transistor case and the heat sink.

The Final Amplifier Section

Phase 3 calls for building a 5-watt power amplifier (PA). Details for this circuit are given in Fig 6-4.

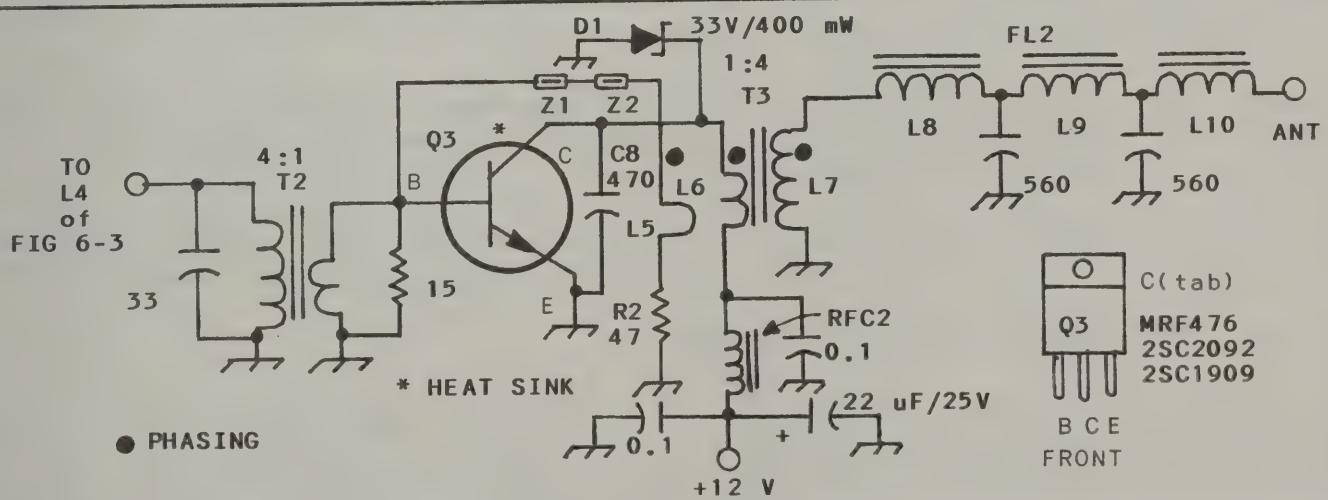


Fig 6-4 -- Circuit for the 5-W class-C amplifier that represents phase 3 of the progressive CW transmitter. Capacitors are disc ceramic except those for FL2, which are silver mica or polystyrene. The polarized capacitor is electrolytic or tantalum. Resistors are 1/4-W carbon film. D1 is a Zener diode. L8 and L9 are 0.8 uH (14 turns of no. 24 enam. wire on an Amidon T50-6 toroid). L9 is 1.67 uH (20 turns of no. 24 enam. wire on a T50-6 toroid core). T2 has 12 primary turns of no. 26 enam. wire on an Amidon FT-37-43 ferrite toroid. Sec. has 6 turns of no. 26 wire. L5 of T3 is a 1-turn, U-shaped link. L6 has 6 turns of no. 26 enam. looped through an Amidon BLN-43-202 binocular core. L7 has 12 turns of no. 26 enam. wire. Dots indicate start of each winding (observe this polarity). RFC2 has 6 turns of No. 22 enam. on an FT-37-43 ferrite toroid. Z1 and Z2 are mini 850-mu ferrite beads.

The amplifier in Fig 6-4 requires approximately 0.4 watt of drive to produce 5 watts of output power. A channel type of heat sink is needed for keeping Q3 cool. I recommend an extruded aluminum heat sink no smaller than 4 square inches with a thickness of no less than 1/8 inch. The bigger the better up to an area of 9 square inches.

L5, R2, Z1 and Z2 form a negative feedback system that levels the amplifier gain and ensures stability. R2 determines the amount of feedback available. The greater the R2 value the lower the feedback power. L5 samples the collector current and diodes Z1 and Z2 act as inductive reactances in the feedback path, particularly at the lower operating frequencies, such as 3.5 and 7 MHz. This broadband amplifier may be used on any band from 160 through 10 meters by merely changing the constants for FL2. Normalized tables for this filter are provided in **The ARRL Handbook**.

D1 protects Q3 from the effects of high SWR, should the load become shorted or open. The diode also protects the transistor from positive voltage spikes that may appear on the +12-V supply line. The Zener diode does not conduct during normal conditions.

All of the leads for L5, L6 and L7 exit from one end of the T3 balun or binocular core. The transformer may thus be mounted vertically on the PC board. Once in place it should be affixed to the board by means of epoxy cement. This will keep the transformer leads from breaking under stress.

This amplifier may be operated linearly (class A or AB) by lifting the secondary return of T2, bypassing it and routing +0.7 to +1 V through the winding to Q3. This bias voltage should be fairly "stiff" (regulated) to preserve the linearity of the amplifier. A heat sink with no less than 9 square inches of flat area is necessary for linear operation.

FL2 also has a low-pass response. The cutoff frequency is 8 MHz. It ensures that all spurious output energy is 40 dB or greater below the peak carrier value. This meets the FCC regulations for HF operation.

R1 in Fig 6-3 should be adjusted for the correct driving power for Q3 of Fig 6-4. Increase the drive level until you obtain 5 watts of output power from Q3.

A 1-Watt CW Transmitter for 10 & 15 Meters

Worldwide communications are possible with very low power if you operate on 21 or 28 MHz when propagation conditions are favorable. A 1-watt rig and a dipole antenna that is high and in the clear can provide remarkable results. The circuit in Fig 6-5 can be built on the universal breadboard if you do not wish to design and etch a PC board for the project. Low-cost parts are used in the circuit. It should take no more than a few hours to assemble the transmitter and have it in operation.

Q1 of Fig 6-5 is an overtone oscillator. It is followed by a broadband (untuned) class-A RF amplifier, Q2. The PA stage, Q3, is an inexpensive plastic-cased audio transistor that can be purchased as surplus for less than \$1. Output from Q3 is filtered by FL1, a five-element low-pass network.

Amplifier Q2 is the only stage keyed in order to minimize the possibility of chirp. Q4 is a PNP dc switch that has waveform shaping components to prevent the signal from being clicky. The leading and trailing edges of the keyed waveform are rounded

off to provide rise and fall times of approximately 5 milliseconds. I measured the backwave (key up) and determined it is 46 dB below the key-down carrier value. Q3 has a gain of +10 dB and its efficiency is 60%. Total key-down current drain for the transmitter at +12.5 V is 180 mA. Power output is 1 watt into a 50-ohm load.

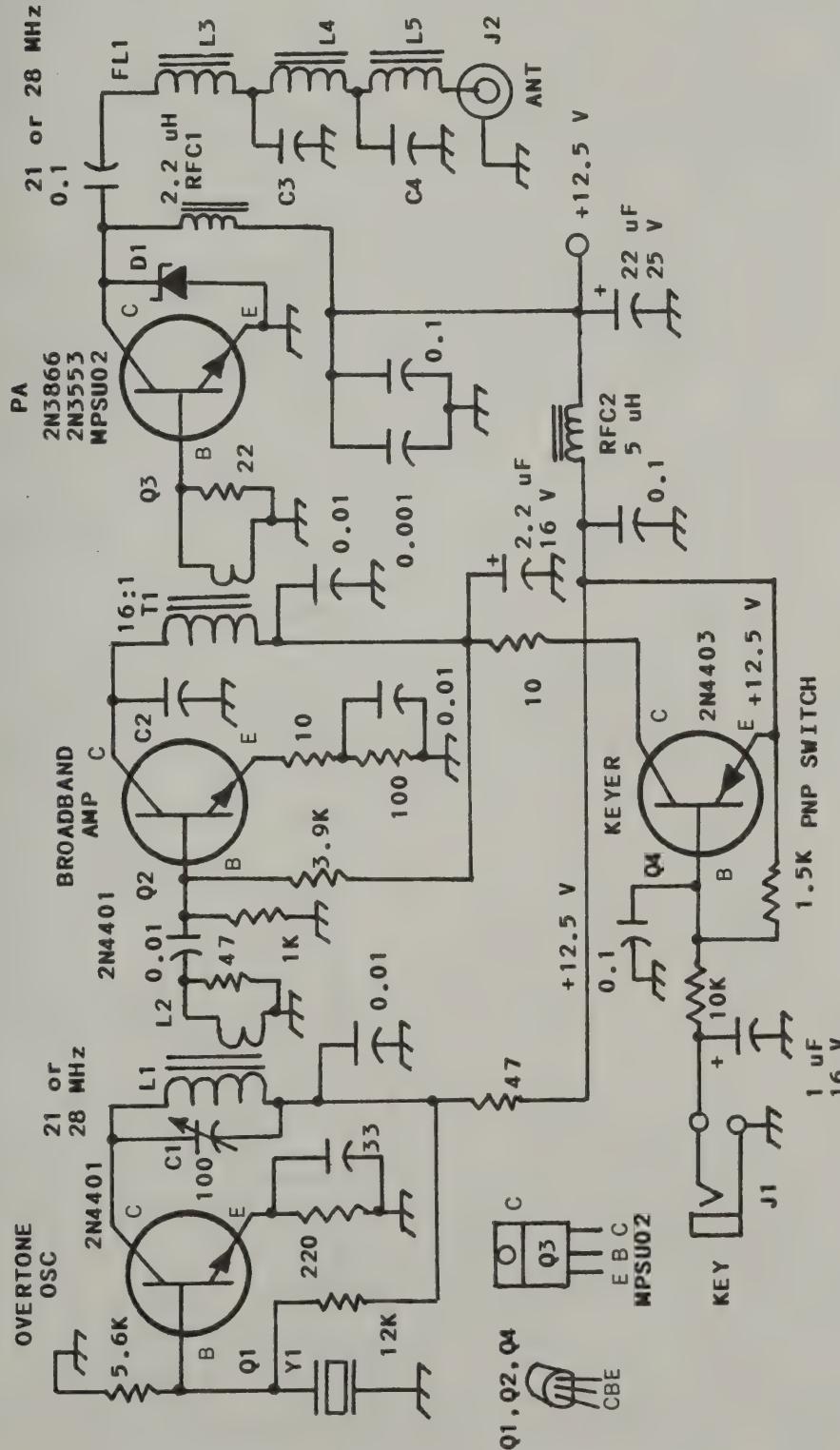


Fig 6-5 -- Schematic diagram of a practical 10- or 15-meter, 1-W CW transmitter. Fixed-value capacitors are disc ceramic unless otherwise specified. Polarized capacitors are tantalum or electrolytic. Resistors are 1/4-W carbon film or composition. C1 is a 10-100 pF ceramic or mica trimmer. C2 = 27 pF for 15m and 12 pF for 10m. C3, C4 = 180 pF for 15m (silver mica) and for 10m, 130 pF. D1 is a 33-V, 400-mW Zener diode. J1 is a two-ckt phone jack and J2 is an RCA phono or SO-239 connector. L1 = 0.8 uH for 15m (14 ts. no 26 enam. on a T50-6 toroid). Link has two turns of no. 26 wire. L1 = 0.48 uH for 10m (10 ts. no 24 enam. wire on a T50-6 toroid). Link has two turns. L3, L5 = 0.26 uH for 15m (8 ts. no. 24 enam. on a T50-6 toroid). L3, L5 for 10m = 0.19 uH (7 ts. no. 24 on a T50-6 core). L4 = 0.55 uH for 15m (12 ts. no. 24 on a T50-6 core). L4 = 0.4 uH for 10m (10 ts. no. 24 on a T50-6 core). T1 = 12 pri. ts. no. 26 enam. on FT-37-43 toroid. Sec = 3 ts no. 26 wire.

Q1 of Fig 6-5 operates on the third overtone of crystal Y1. C1 and L1 form a tuned circuit at the crystal overtone frequency. C1 is adjusted for reliable oscillation of Q1. Generally, this requires that the tuned circuit be adjusted for resonance slightly above the overtone frequency. No other transmitter tuning is required.

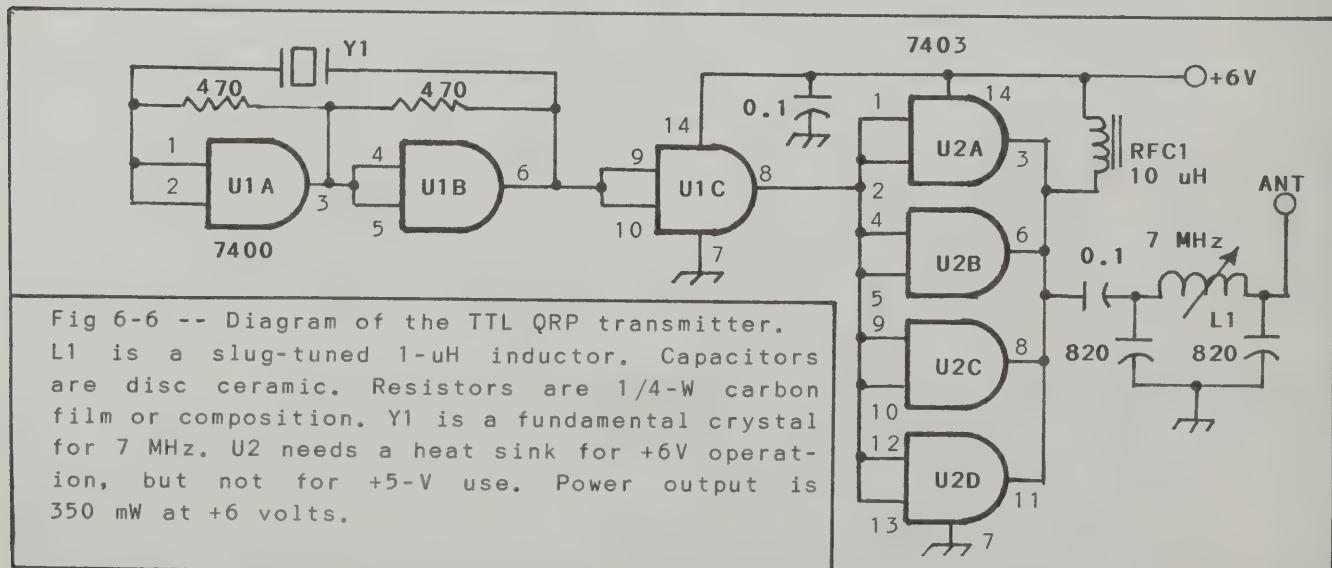
Various transistors may be used in this circuit. There is no reason why you can't use a 2N2222A or 2N4400 for Q1 and Q2, although the 2N4401 will provide slightly more gain. A 2N5179 or 2N5770 may also be used. A 2SC779 is a suitable substitute for the devices listed above Q3.

Y1 is not especially critical provided it is a plated 3rd-overtone crystal. The load capacitance may be 20 or 30 pF. I prefer crystals that are in HC-6/U style holders rather than the smaller HC-18/U types. I do not recommend that you try to use surplus FT-243 crystals on their overtone mode. Sluggish performance and low output from Q1 may result from this practice.

There is no reason why you can't add the amplifier in Fig 6-4 if you want to boost the power to 5 watts. The only changes required are for the components in FL2. The Fig 6-5 FL1 values may be used for the 5-W amplifier. Do not bunch the filter components close together. It is better to keep the toroids separated by at least 1/2 inch and in a straight line. This will help ensure good input-output isolation for FL1, and hence better harmonic suppression. Too close a spacing leads to unwanted stray coupling between the elements of the filter. Make certain that the antenna system presents a 50-ohm nonreactive load to the Fig 6-5 transmitter. Filters such as FL1 are designed for a specific impedance (50 ohms in this example), and a mismatch can spoil the filter performance.

Experimental TTL IC Transmitter

The Spring 1990 edition of **SPRAT**, the official journal of the G-QRP Club, carried a cute circuit for a simple CW transceiver that uses two ICs and a pair of diodes. It was designed by WB0NQM for operation on 40 meters. The author states that it delivers 360 mW of output power when operated from a +6-V supply. The transmitter circuit is worth presenting here, should you wish to experiment with TTL chips for RF service. Fig 6-6 shows the circuit.



I want to stress that I have not tried the circuit in Fig 6-6. Were I to put it on the air I would definitely add a harmonic filter after the pi network. I presume that the circuit can be keyed by breaking the +6-V lead to pin 14 of U1. You may wish to key the U1 pin 7 lead instead. Certainly, this little transmitter could be used nicely as a miniature emergency rig for hikers and campers.

A Simple 1.5 Watter for 30 Meters

Fig 6-7 is the circuit for a practical CW transmitter that uses only two transistors. It delivers 1.5 watts of output power when a +12.5-V dc supply is used. The circuit is a variation of the popular "Cubic Inch" that was presented in July 1982 *QST* by D. Monticelli, AE6C. Information concerning 40- and 80-meter operation may be obtained from the 1990 *ARRL Handbook*, page 30-41.

My version includes CW waveform shaping via a PNP keying switch. I also included a half-wave harmonic filter at the transmitter output. All spurious energy is 40 dB or greater below peak carrier value. The keying transistor may be eliminated if you key the +12.5-V line to Q1. The CW note will be somewhat hard (clicky) without the keying transistor, Q2. I have no reason to doubt that this circuit can be used successfully on 20 meters if appropriate changes are made to the tuned circuit, the values of C1 and C2 and FL1.

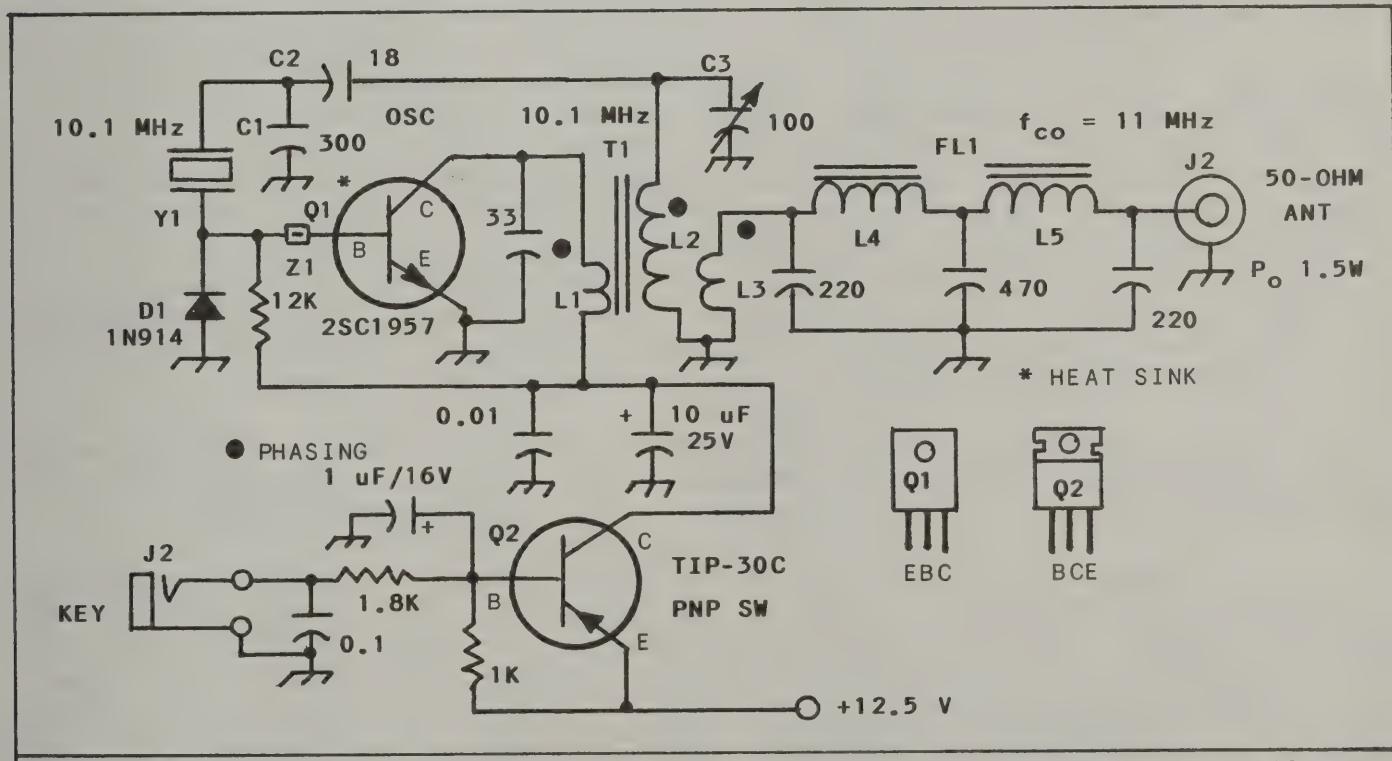


Fig 6-7 -- Practical 30-meter CW transmitter. FL1 capacitors are silver mica or polystyrene. Others are disc ceramic. Capacitors with polarity marked are tantalum or electrolytic. Resistors are 1/4-W carbon film or composition. C3 is a 10-100 pF ceramic or mica trimmer. L2 of T1 has 26 ts. of no. 26 enam. wire on an Amidon T50-6 toroid. L1 has 3 ts. of no. 26 wire and L3 has 4 ts. of no. 26 wire (observe polarity). L4 and L5 have 18 ts. of no. 26 enam. wire on a T37-6 toroid. Y1 is a fundamental crystal, 30 pF load capacitance.

Various types of transistors may be used at Q1 and Q2 provided they have similar electrical characteristics. An MRF472 is a good substitute for Q1. Likewise for a 2SC799. The 2SC devices are available from ORA Electronics in Chatsworth, CA 91313 and from Oak Hills Research, 20879 Madison, Big Rapids, MI 49307. Any PNP transistor that can handle 0.5-A or greater may be used for Q2.

C1 and C2 in Fig 6-7 are feedback capacitors. The C2 value may be increased if your crystal does not oscillate when T1 is tuned to the operating frequency. Adjust C3 for a chirp-free CW note, consistent with optimum RF output power. Adjustment of C3 will cause some change in the operating frequency. Z1 is a mini ferrite bead (850 μ u). It helps prevent VHF self-oscillation. Likewise for the 33-pF capacitor from the Q1 collector to ground.

Q1 draws 100 mA (key down) when Y1 is removed. Operational key-down current is 200-250 mA. The transmitter efficiency is roughly 52%. Output power is 1.75 W when a 14-V power supply is used. The waveform has a 5 ms rise and fall time.

A small heat sink is required for Q1. I used a 1-1/2 inch piece of hardware store aluminum angle stock. The material is 1/8 inch thick. No heat sink is needed for Q2. My transmitter is built on one of the universal PC boards described earlier in this book. (available from FAR Circuits).

A 10-W Linear Amplifier for CW or SSB

Although it is possible to use bipolar transistors or power FETs in parallel to obtain increased RF output power, a better scheme is to use two devices in push pull. This makes the individual collector impedances higher, and therefore easier to match to the load. Also, the push-pull arrangement tends to cancel even harmonics of the operating frequency. Fig 6-8 shows the circuit for a practical 10-W linear solid-state amplifier. It may be used from 1.8 to 30 MHz by only changing the FL1 circuit constants for the band of choice. A group of harmonic filters may be band switched at the output of T2 to provide multiband operation. The switching process is uncomplicated, owing to the 50-ohm terminal impedance of the filters. A two pole, multiposition rotary wafer switch may be used for this purpose.

T1 and T2 in Fig 6-8 are broadband transformers that are wound on 850- μ u ferrite cores (Amidon no. 43 material). T1 matches the 50-ohm source to the bases of Q1 and Q2. T2 matches the Q1 and Q2 collectors to the 50-ohm harmonic filter.

C1 is a reactance compensation capacitor that helps ensure a low SWR between the signal source and the amplifier. C2, C3, R1 and R4 are used in gain-leveling networks. In effect, the amplifier receives maximum drive at the upper end of the HF spectrum, whereas the driving power declines (at the Q1, Q2 bases) as the operating frequency in MHz is lowered. This helps to assure uniform amplifier output power across the MF and HF spectrum. R2 and R3 are used to de-Q the amplifier input circuit for stability enhancement. They help to lower the Q of RF chokes RFC1 and RFC2.

In order to establish linearity it is necessary to apply forward bias to the amplifier transistors. This is accomplished by way of D1. It sets the base bias at +0.7 V and causes a resting collector current to flow for class AB operation. C8 charges and helps to stabilize the forward bias.

Negative feedback is provided by means of R6, Z1 and Z2. The feedback also helps to equalize the amplifier gain from 1.8 to 30 MHz, while aiding the stability.

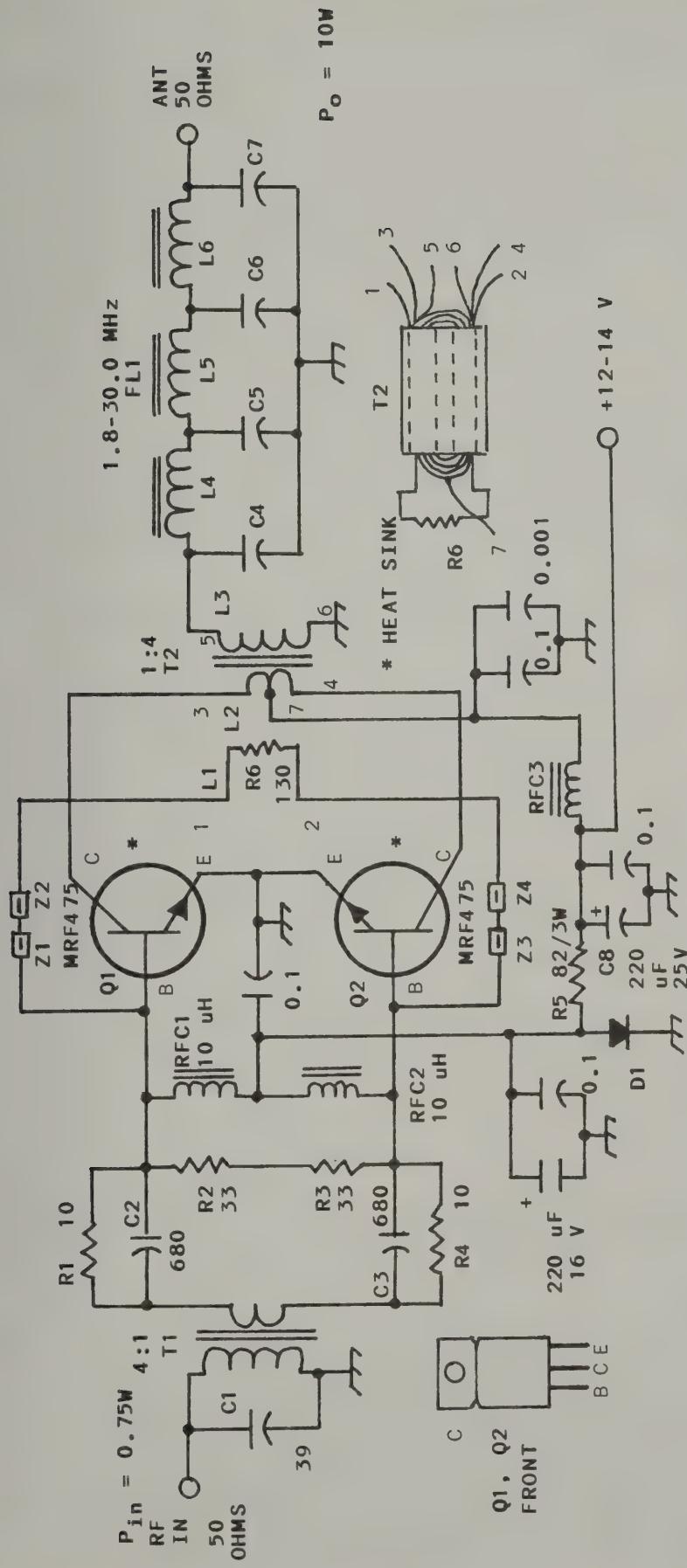


Fig 6-8 -- Schematic diagram of a practical 10-W linear amplifier. Capacitors, other than C_4 , C_5 and C_6 , are disc ceramic. C_4 , C_5 and C_6 are silver mica or polystyrene. Polarized capacitors are tantalum or electrolytic. L_1 of T_2 is a U-shaped wire passed through an Amidon BLN-43-7051 balun core (850 μ u). L_2 is a U-shaped, one-turn piece of no. 22 enam. wire through the two core holes, center tapped at bend in U. L_3 has two turns of no. 22 enam. or hookup wire through balun core. See Table 6-1 for data about FL_1 . Resistors are 1/2-W carbon film or composition unless otherwise indicated. RFC_1 and RFC_2 are mini 10-uH RF chokes. RFC_3 has 6 turns of no. 22 enam. wire on an FT-50-43 ferrite toroid. T_1 primary has 10 turns of no. 24 enam. wire on an FT-50-43 ferrite toroid (850 μ u). Secondary has 5 turns of no. 24 enam. wire. Z_1 through Z_4 , inc., are Amidon FB-43-301 ferrite beads.

TABLE 6-1

BAND	C4, C7 (pF)	C5, C6 (pF)	L4, L6 (uH)	L5 (uH)
80	510	1300	2.64, 21 ts. no. 24 on T68-2 toroid.	3.26, 24 ts no. 24 on T68-2 toroid.
40	330	750	1.5, 16 ts. no. 24 on a T68-2 toroid.	1.79, 18 ts. no. 24 on T68-2 toroid.
30	180	470	0.95, 14 ts. no. 24 on T68-6 core.	1.19, 16 ts. no. 24 on T68-6 toroid.
20	180	390	0.77, 13 ts. no. 24 on a T68-6 toroid.	0.90, 14 ts. no. 24 on a T68-6 toroid.
15	130	270	0.53, 11 ts. no. 24 on a T68-6 toroid.	0.6, 11 ts no. 24 on a T68-6 toroid.
10	82	180	0.36, 9 ts. no. 24 on a T68-6 toroid.	0.42, 10 ts. no 24 on a T68-6 toroid.
Component values for FL1 of Fig 6-8. These 7-element low-pass filter constants were developed around standard-value capacitors by E. Wetherhold, W3NQN. Filter cutoff frequencies from 80 through 10 meters, respectively, in MHz, are 3.81, 7.23, 10.33, 14.4, 21.48 and 30.9. Values for 12, 17 and 160 meters may be obtained from the normalized filter tables in The ARRL Handbook .				

The Fig 6-8 amplifier may be built on double-sided PC-board stock. Isolated pads can be etched on one surface of the board, or they may be established by way of a hobby motor with a small cone-shaped abrasive bit. All of the components can be mounted on the etched side of the board.

Q1 and Q2 require a fairly large heat sink for linear service. I recommend that you use an extruded aluminum sink that has an area no less than 16 square inches. The TO-220 transistors must be insulated from the heat sink by means of mica wafers and nylon insulating washers. Heat-sink compound is necessary between the transistors and the heat sink.

Q1 and Q2 are Motorola Citizen's Band transistors. They are capable of delivering in excess of 5 watts of output apiece. Power output from this circuit may reach 15 watts with a 14.6-V dc supply. Efficiency is approximately 55%. Minimum driving power (P_{in}) is 0.75 W. Full output power will occur with 1.0 to 1.5 watts of excitation power into T1. A regulated +12-volt power supply that can deliver 1.75 A or greater is required for this amplifier. D1, a 2-A, 50-PIV rectifier diode, draws part of the overall amplifier current. All spurious output from the Fig 6-8 amplifier is 40 dB or greater below peak output power.

Detailed information about how to design an SSB transmitter, along with practical circuits, was presented in **QST** by myself. See Sept., Oct., Nov. and Dec. of 1985 **QST**. The final installment of "Principles and Building of SSB Gear" was published in Jan. 1986 **QST**.

Complete information and PC-board patterns for a 60-watt solid-state amplifier are available from Motorola, P.O. Box 20912, Phoenix, AZ 85036. The Engineering bulletin is entitled "60 Watt VHF Amplifier," no. EB-93. A 1.8 to 30 MHz 600-W power FET linear amplifier is treated in detail in Motorola EB-104. Also, numerous practical circuits and PC layouts for HF-band amplifiers are found in the **Motorola RF Device Data book**, no. DL110.

25 Watt Power FET Amplifier

There are many power FETs that were not designed for RF service. These devices are intended primarily for ac and dc switching and control service. A number of them work well at MF and HF. Two low-cost IRF511 switching FETs are used in this circuit. They will deliver up to 25 watts of RF output after filtering when used in parallel. Fig 6-9 illustrates the circuit.

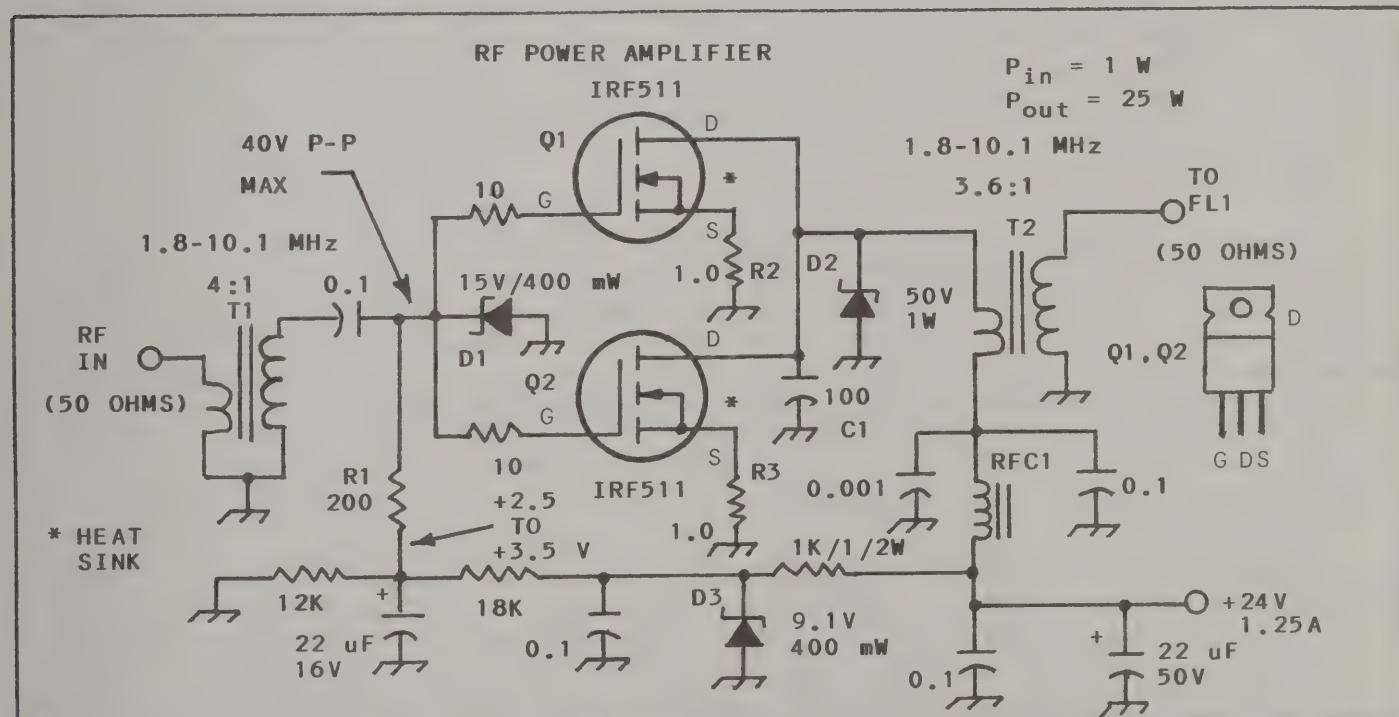


Fig 6-9 -- Circuit for a practical 25-W FET amplifier. Capacitors are disc ceramic. Polarized units are tantalum or electrolytic. Resistors are 1/4- or 1/2-W carbon film or composition. D1, D2 and D3 are Zener diodes. RFC1 has 8 turns of no. 22 enam. wire on an FT-50-43 toroid. T1 has 12 sec. turns of no. 26 enam. wire on an FT-50-43 toroid. Pri. has 6 turns of no. 26 to cover all of the sec. winding. T2 pri, has 2 turns of no. 24 hookup wire on an Amidon BN-43-3312 balun core. Sec. has 4 turns of no. 24 enam. wire. Q1 and Q2 are insulated from the heat sink, or the sink must be insulated from the PC board ground bus. Heat-sink area is 12 square inches or greater.

You may use the filter constants in Table 6-1 for the FET amplifier in Fig 6-9. The normalized low-pass filter tables that are presented in *The ARRL Handbook* may be consulted, should you wish to design your own filters.

The two 10-ohm gate resistors are for lowering the input-circuit Q and to serve as VHF parasitic suppressors. They need to be placed as close to the Q1, Q2 gate pins as possible. D1 protects the delicate transistor gates from excess P-P voltage (positive excursions). R1 establishes the amplifier input impedance at 200 ohms. R2 and R3 serve as ballast resistors which tend to balance the currents of the two transistors so that one won't draw more current (current hogging) than the other when the transistor gains aren't closely matched.

D2 is a back-up protection Zener diode. Q1 and Q2 have built in 50-V Zener diodes from drain to source. This diode is probably not necessary. The forward gate bias may be changed in accordance with the class of service you desire for this amplifier. I use +2 V for CW operation and 3.5 V for linear service. The resting drain current (no signal) for linear amplification should be 100-150 mA for the pair of IRF511s. The amplifier in Fig 6-9 has an efficiency of 70%.

The performance of these particular FETs is poor above 10.1 MHz. Efficiency drops quickly at 14 MHz, and declines as the operating frequency is increased. N7RK at Motorola suggests the MTP3055E as a more rugged substitute for the IRF511. The maximum power-dissipation rating for this transistor is 40 W and the RDS(on) is a mere 0.15 ohm, which suggests that it should be an excellent performer at HF. It is also a low-cost switching FET. Power FETs that are designed for RF service at HF and VHF are very costly.

Unlike bipolar RF power devices, power FETs do not operate efficiently at reduced drain-source voltage. For example, an IRF511 delivers a maximum output power of roughly 5 watts at +13.6 V, whereas it can yield up to 15 watts of output power at +24 V. The lower operating voltages cause the device to saturate quickly (no further increase in output power) even though the same amount of driving power is applied. Increasing the forward bias, and hence the resting drain current, will increase the output power, but the efficiency will be very poor.

A Switch-Around 15-W, 2-Meter Amplifier

It is convenient and sometimes necessary to elevate the output power of a 2-meter FM handi-talkie from one or two watts to 10 or more watts. This can be done easily with a single VHF power transistor such as the Motorola MRF262. A class-C amplifier is suitable for CW and FM amplification, but not for SSB operation. The circuit in Fig 6-10 is a practical example of such an amplifier. A less elaborate version of this amplifier was presented in *QST* for August 1984 by W1FB ("Some Basics of VHF Design and Layout").

It is necessary to employ an RF-sensed TR switch when you operate a handi-talkie into an external amplifier. This feature is included in the Fig 6-10 circuit. Also, effective harmonic filtering is used at the amplifier output to ensure purity of emissions in accordance with FCC regulations. Photographs of the amplifier are provided in the August 1984 issue of *QST*.

The amplifier may be driven to 10-W output with an RF input power of 1 W. A 2-W handi-talkie will increase the amplifier output power to 15 W.

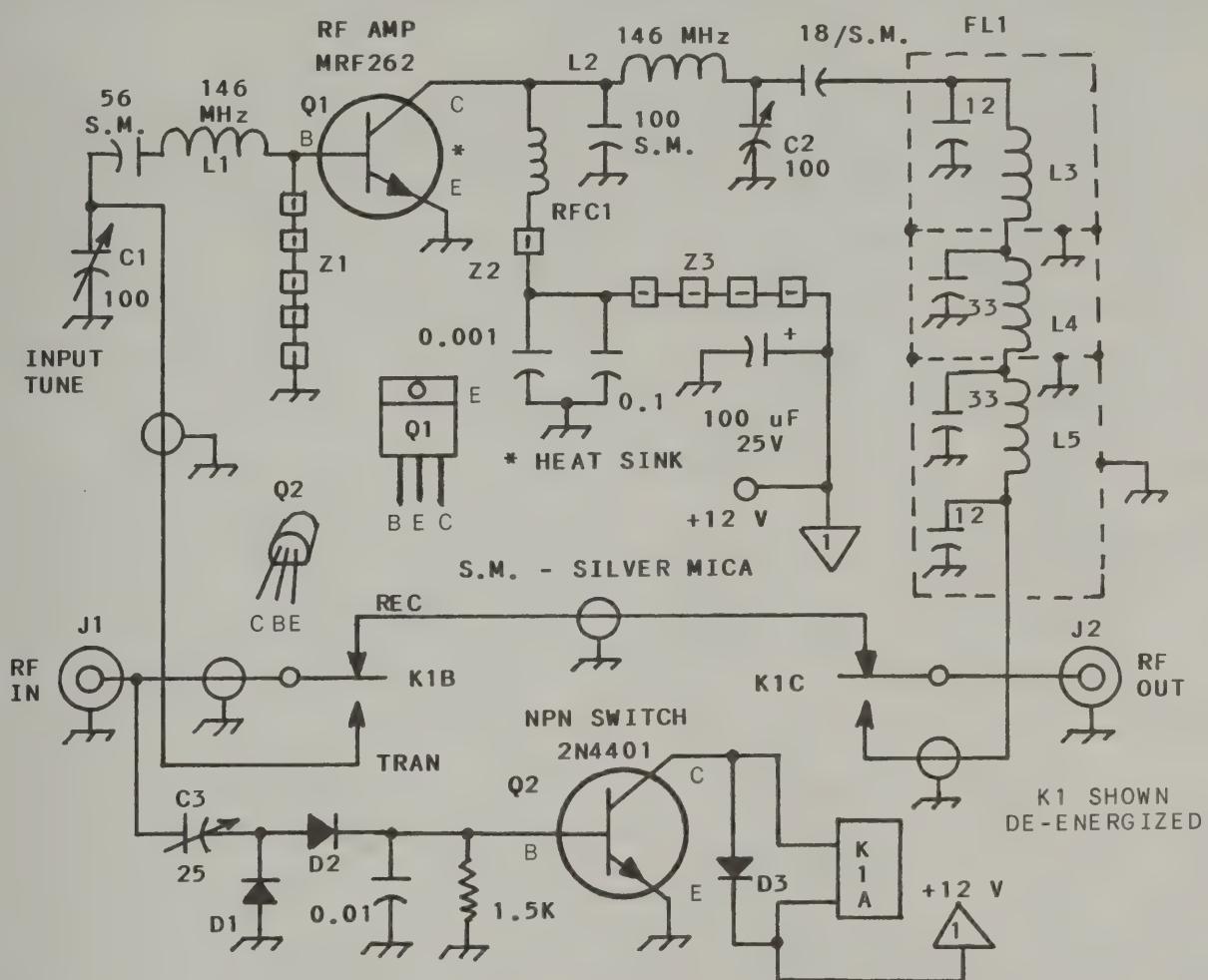


Fig 6-10 -- Practical 2-meter switch-through 15-W amplifier. Fixed-value capacitors are disc ceramic unless otherwise noted. All FL1 capacitors are silver mica. Polarized capacitor is tantalum or electrolytic. The resistor is a 1/4-W carbon film unit. C1 and C2 are 10-100 pF ceramic or mica compression. C3 is a 2-25 pF ceramic or plastic trimmer. D1, D2 and D3 are 1N914s. J1 and J2 are type BNC or SO-239, chassis mount. K1 is a DPDT 12-V dc mini relay. L1 and L2 have two turns of no. 14 wire, 5/16" ID by 3/8" long. L3, L4 and L5 have four turns of no. 20 enam. or bus wire, 5/16" ID by 3/8" long. Shields are used between L3, L4 and L5. Z1 uses five mini 40-mu Amidon ferrite beads (no. 64 material) on a piece of bus wire. Z2 is a single 850-mu mini bead (no. 43 material). Z3 consists of four mini beads, 850 mu, no. 43 mix. RFC1 has 13 close-wound turns of no. 24 enam. wire on the body of a 1/2-W, 5.6K-ohm carbon composition resistor. The ends of the choke winding are common to the resistor pigtails.

RG-174 coaxial cable is used for the signal leads in the above circuit. The shield braid of each coaxial line should be grounded at each end. A heat sink with no less than 9 square inches of flat area, 1/8 inch or greater thickness, is required for Q1 of Fig 6-10. All signal leads must be as short and direct as you can make them. The PC-board conductors should be wide in order to minimize unwanted stray inductance. A PC-board etching template and parts-placement guide for the basic

amplifier was published in the W1FB August 1984 **QST** article. FL1 may be built in a small shield box made from double-sided PC-board sections. The relay sensing and control circuit can be assembled on a small piece of perf board or regular PC board.

Adjust C3 for maximum capacitance before testing this amplifier. Next, apply 1-watt of excitation while a 50-ohm dummy antenna is connected to J2. Adjust C1 and C2 for maximum output power as indicated by an RF wattmeter or RF probe. Adjust these trimmers a second time to compensate for any interaction that may occur. Finally, adjust C3 for the least capacitance that will ensure reliable triggering of K1. This adjustment will need to be repeated if you increase the driving power to two watts. Likewise with the settings of C1 and C2.

Transistors other than the MRF262 may be used at Q1 if they have similar characteristics. A 2N2222A or a 2N4400 may be used for Q2.

A 1.5-Watt 40-Meter CW Transceiver

It is convenient and fun to carry with you a small, low-power transceiver when on vacation, camping or hiking. I developed the circuit in Fig 6-11 for those times when I am afield and have access to a 12-V vehicular or boat battery. You will note the relative simplicity of the circuit. The low parts count and cost results from the use of a DC receiver and crystal control.

The Receiver

A doubly balanced product detector is used at U1. This popular NE602 provides high conversion gain and by virtue of being a doubly balanced detector it does a nice job of rejecting AM signals that generally blanket DC receivers that have such detectors as the 40673 or similar dual-gate MOSFETs. A low-noise, high-gain audio preamp follows the detector (Q1). A 741 op amp, U2, provides more than ample audio gain for driving a pair of 8-ohm headphones. Transmitting crystal Y1 is "rubbered" during receive to provide RIT action when Q2 is used for injecting the U1 detector. RIT (VXO) control is by means of C21, a 100-pF mini tuning capacitor. The local oscillator, Q2, is switched for non-VXO use during transmit by D9 and D10. A sidetone signal for monitoring your sending is routed to U2 during transmit.

Transmitter Circuit

The local oscillator output is routed to Q3 via tuned transformer T3. You may substitute a 2N3553 transistor for the NEC 2SC799 device shown in Fig 6-11. A heavy-duty aluminum press-on heat sink (TO-5) is required if you use a 2N3553. Output from Q3 is filtered by FL1. Q2 and the sidetone oscillator (Q5 and Q5) are keyed during transmit. SWR protection for the final amplifier, Q3, is ensured by Zener diode D5. D4 protects the transceiver from a cross-polarized connection to the power supply.

Transceiver Performance

The receiver is sensitive enough to make a 0.1-uV signal just discernible above the receiver noise floor. During testing I actually heard some CW signals with the antenna disconnected from the transceiver. Touching my finger to front-end protection diodes D1 and D2 caused these signals to be quite loud.

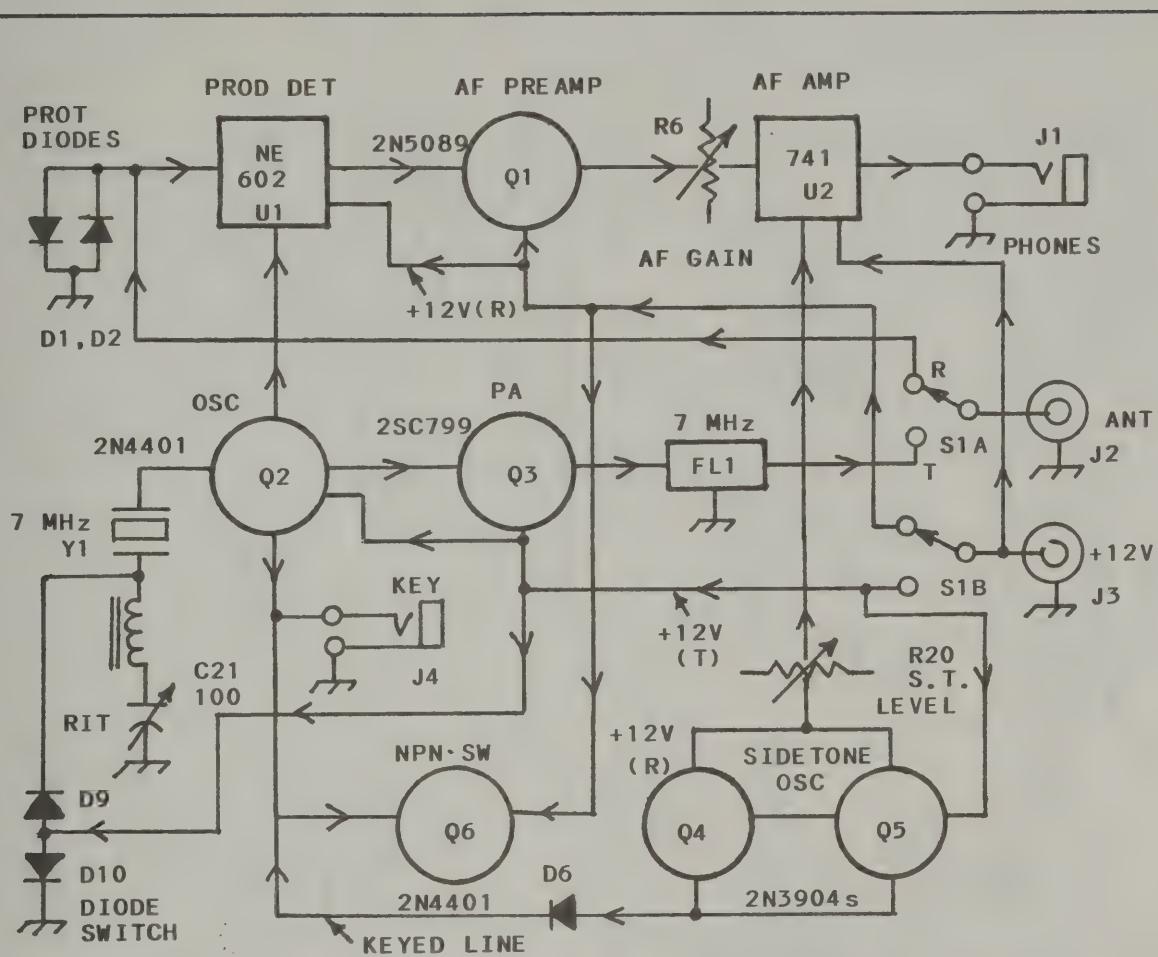


Fig 6-11 -- Block diagram of the 40-meter CW transceiver. The arrows indicate the direction of signal and dc current flow. Q2 and Y1 operate as a VXO during receive to provide the RIT function. D9 and D10 disable the VXO circuit during transmit to allow Q2 to operate as a straight crystal-controlled oscillator. The NPN dc switch, Q6, shorts out the keying line during receive to allow Q2 to oscillate continuously. A diode gate (D6) prevents the sidetone oscillator from functioning when Q6 is in the ON state.

The block diagram above shows how the transceiver circuits mate with one another. TR switching is done by means of S1, a DPDT toggle switch. Diode and bipolar-transistor switching is initiated also when S1 is cycled from transmit to receive.

Fig 6-12 shows the receiver portion of the circuit. This part of the transceiver may be built as a separate receiver if you wish. A tunable oscillator (VFO) can be substituted for the crystal oscillator in Fig 6-13 to provide full coverage of the 40-meter band, or a different band of your choice. The only circuit changes necessary for receiver operation on 160, 80, 30 or 20 meters involve the input tuned circuit (C1 and T1) and the proper choice of VFO operating frequency.

The 2N5089 used at Q1 is a low noise, high gain audio transistor. I chose it in the interest of minimum receiver noise. There is no reason why you can't substitute a 2N3904 or similar transistor for the device specified. A further reduction of receiver noise (hiss) may be possible by using a biFET op-amp at U2.

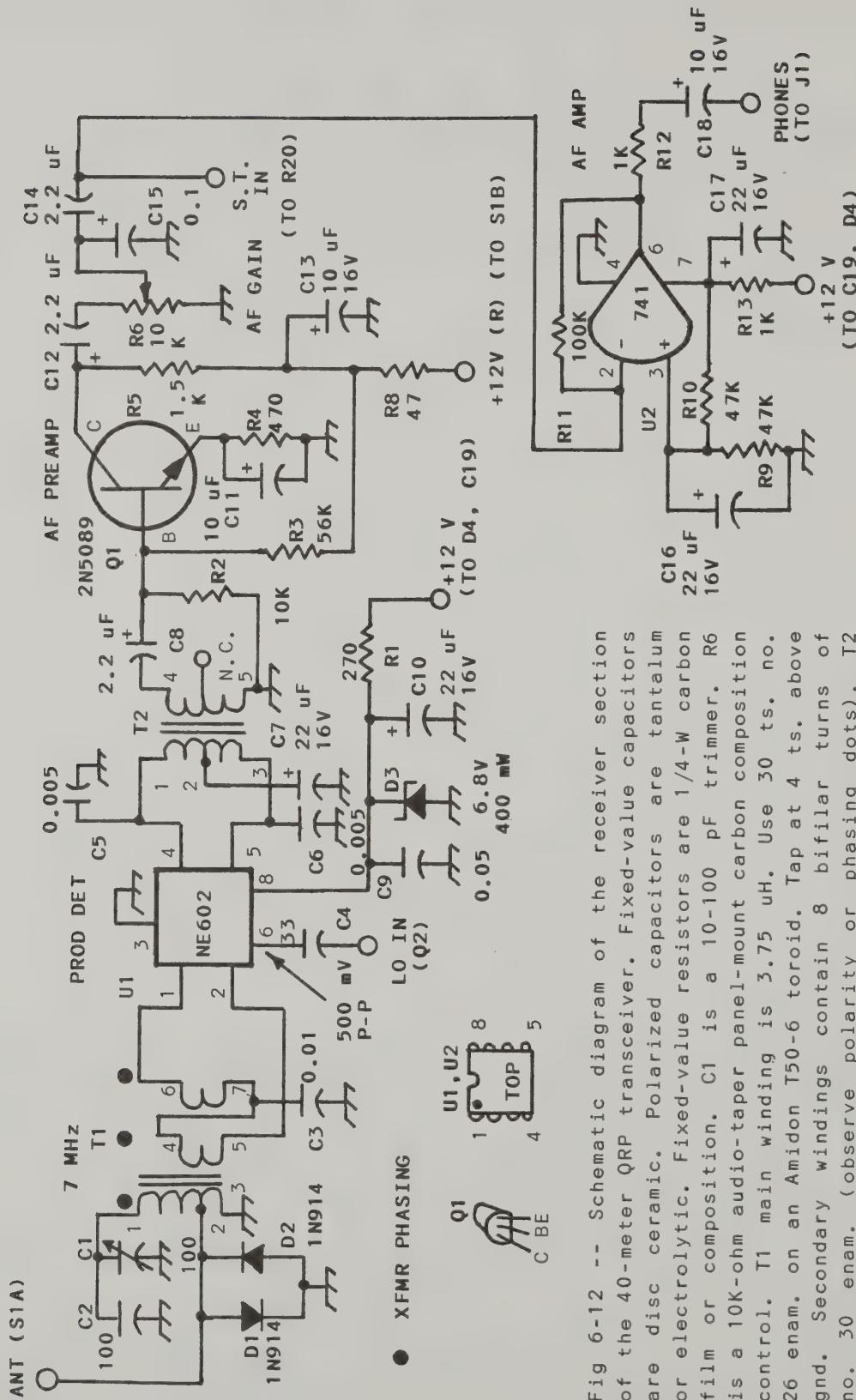


Fig. 6-12 -- Schematic diagram of the receiver section of the 40-meter QRP transceiver. Fixed-value capacitors are disc ceramic. Polarized capacitors are 1/4-W carbon film or composition. Fixed-value resistors are 10-100 pF trimmer. R6 is a 10K-ohm audio-taper panel-mount carbon composition control. T1 main winding is 3.75 uH. Use 30 ts. no. 26 enam. on an Amidon T50-6 toroid. Tap at 4 ts. above gnd. Secondary windings contain 8 bifilar turns of no. 30 enam. (observe polarity or phasing dots). T2 is a Mouser no. 42TL021 4K ohm c.t. to 600 ohm c.t. PC-mount audio transformer or equivalent. Secondary center tap not used. C5 and C6 tune the T1 primary to 700 Hz to aid CW selectivity. C15 rolls off the high-frequency response.

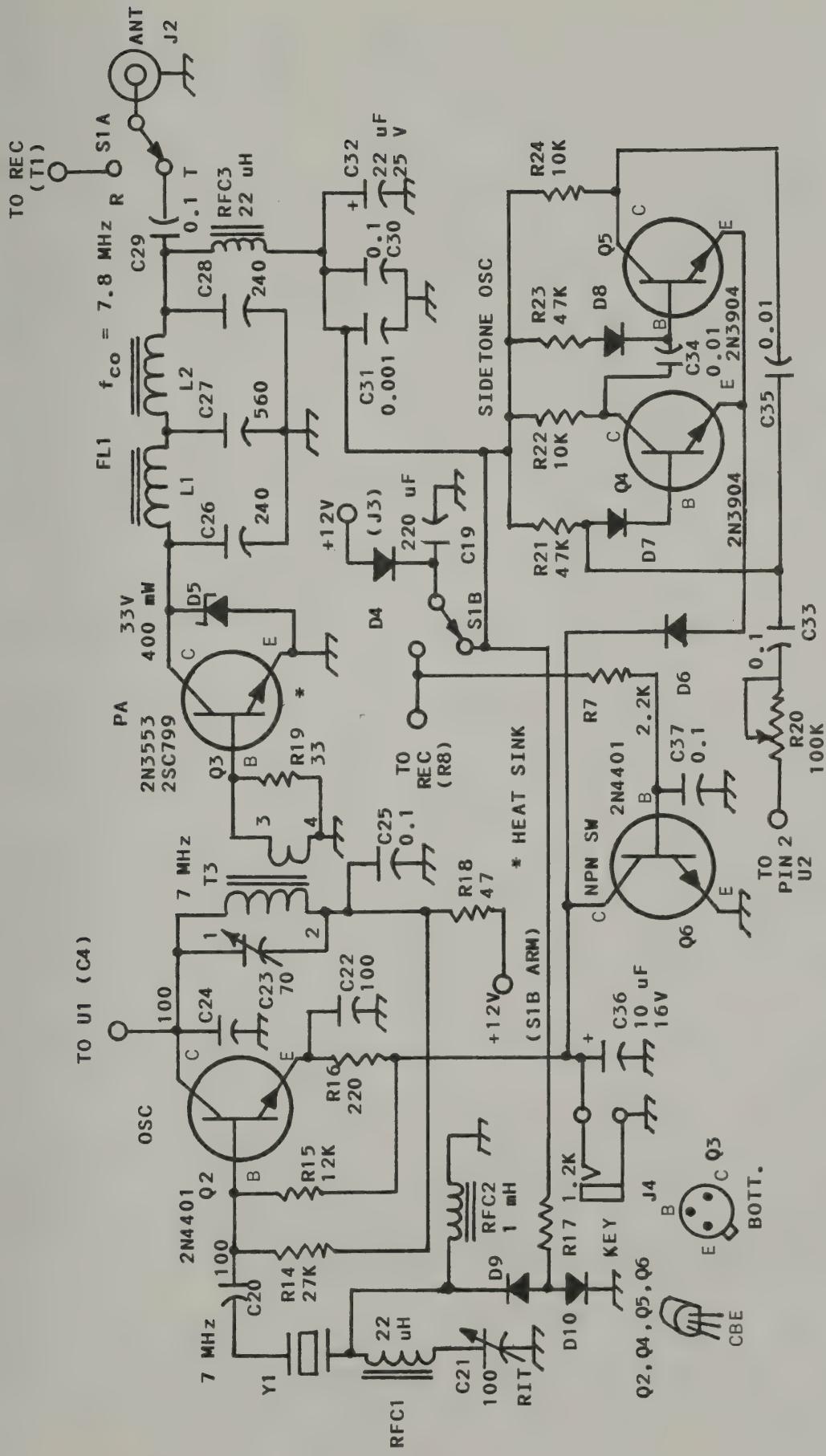


Fig 6-13 -- Circuit for the transmitter and sidetone oscillator. Components and values are as follows:

- OSC:** 2N4401, Y1, Y2, C20, C21, R14, 27K, R15, 12K, 220, C22, 100, C23, 70, T3, 7 MHz, 400 mW, 33V, 2N3553, PA, 33V, 400 mW, 2N3799.
- Driver:** T1, 2N3799, C24, 100, C25, 0.1, T2, 1.12 uH (17 ts no. 26 enam. on T37-2 toroid).
- Power Amplifier:** T3, 2N3553, C26, 1.2, C27, 240, C28, 560, C29, 0.1, T1, 1.12 uH (17 ts no. 26 enam. on T37-2 toroid).
- Side Tone Oscillator:** 2N3904, C31, 0.001, C32, 22 uF, C33, 0.1, C34, 0.01, C35, 0.01, C36, 10 uF, 16V, Q6, 2N4401, R7, 2.2K, D6, D7, D8, D9, R16, 220, R18, 4.7, R21, 4.7K, R22, 10K, R23, 4.7K, R24, 10K.
- Control:** S1A, S1B, S1C, KEY, CBE, BOTT.
- Antenna:** ANT, J2, J1.

Capacitors and resistors are the same as for Fig 6-12. C21 is a 10-100 pF mini air variable (panel mount). C23 is a mini 70-pF plastic trimmer. C26, 27 & 28 are silver mica or poly. D6-D10 are 1N914. D4 is a 50 PRV, 1-A rect. diode. L1, L2 = 1.12 uH (17 ts no. 26 enam. on T37-2 toroid). RFC1, 2 & 3 are mini RF chokes (Mouser type 43LR). R20 is a mini PC-mount carbon control. S1 is a mini DPDT toggle. T1 pri has 28 ts no. 26 enam. on a T50-2 toroid. Sec has 5 ts no. 26 over C25 end of primary. Y1 is a fundamental general-purpose crystal, 30-pF load capacitance (panel mount in socket).

The transmitter and sidetone section of the transceiver is shown in Fig 6-13. When D9 and D10 are in the OFF state the oscillator becomes a VXO by virtue of C21 and RFC1. These reactances in the crystal ground path make it possible to shift an AT-cut plated crystal some 7 kHz at 40 meters. The diodes are switched on during transmit via S1B, at which time they conduct and short out C21 and RFC1. Y1 should be capable of being shifted approximately 1.5 kHz above the marked value when the VXO is engaged. This allows the circuit to be used as an RIT during receive -- a needed feature when someone answers your CQ and is not exactly on your frequency.

The LO, Q2, is emitter keyed during transmit. The sidetone oscillator (Q4 and Q5) is keyed along with Q2. NPN switch Q6 is turned on by S1B during receive to complete the emitter circuit of the oscillator. Gating diode D6 prevents the sidetone oscillator from receiving + voltage from the emitter of Q2 during key-up. C36 rounds off the sharp edges of the keyed waveform to prevent clicks.

R19 is a de-Qing base resistor to enhance the Q3 stability. D5 protects Q3 from SWR damage or spikes that may appear on the +12-V supply line. It does not conduct until the peak collector voltage reaches 33. Normal peak collector voltage is 24. FL1 suppresses harmonic currents and matches the Q3 collector to a 50-ohm antenna load.

Q4 and Q5 form a multivibrator for which the operating frequency is approximately 700 Hz. Output from the sidetone oscillator is routed to the 741 op amp, U2, to provide a 700-Hz tone for monitoring your sending. R20 is adjusted for comfortable headphone volume while listening to the sidetone.

Power output from the transmitter is 1.5 watts when a 12.5-V power supply is used. There is a 0.7-V drop across polarity-guarding diode D4. Total transceiver dc current with the key down is 190 mA. At +13 V the current increases to 210 mA and the output power rises to 1.75 W. The transceiver draws roughly 55 mA during receive.

You may want to build only the transmitter portion of this transceiver. It may be used on 160, 80, 30 or 20 meters by changing the Q2 tuned circuit (C2 and T3) accordingly. It should resonate at the crystal frequency. FL1 will also need to be modified for the band of operation. You may use the normalized filter tables in **The ARRL Handbook** to obtain the new FL1 values. RFC1 should be increased to 100 uH for 160 and 80 meters. Use a 15-uH RF choke for 20-meter operation.

R19 may require changing for various other frequencies. I suggest 10 ohms for 160 and 80 meters, and 47 ohms for 20 meters. The indicated value is okay for 40-meter operation.

C22 at Q2 acts as part of a feedback network in combination with the internal capacitance of the transistor. The C22 value will need to be increased to 220 pF for 80 meters and 470 pF for 160 meters. A 56-pF value should be correct for 20 meters. Experiment with the C22 value until you obtain reliable oscillator starting and chirpless keying. The tuning of C23 also affects the keyed waveform. It should be set for a chirp-free note, consistent with maximum Q3 power output.

A number of crystals may be used at Y1 if you add a crystal-selector switch. The leads from the switch to the crystals and PC board should be short and direct if this is done. The lower the operating frequency (80 meters, for example) the less frequency shift for Y1. A typical VXO shift on 3.5 MHz is 1.5 kHz. At 20 meters it may be as great as 15 kHz.

a - S1A (ANT), b - TO J4 (KEY), c - TO S1B (REC).
 d - AF OUT, J1, e - TO R6 HIGH, f - TO Y1, g - TO J4 (KEY).
 f - CKT GND, i - TO Y1, j - TO S1B (TRAN)

VERTICAL MT. RESISTOR

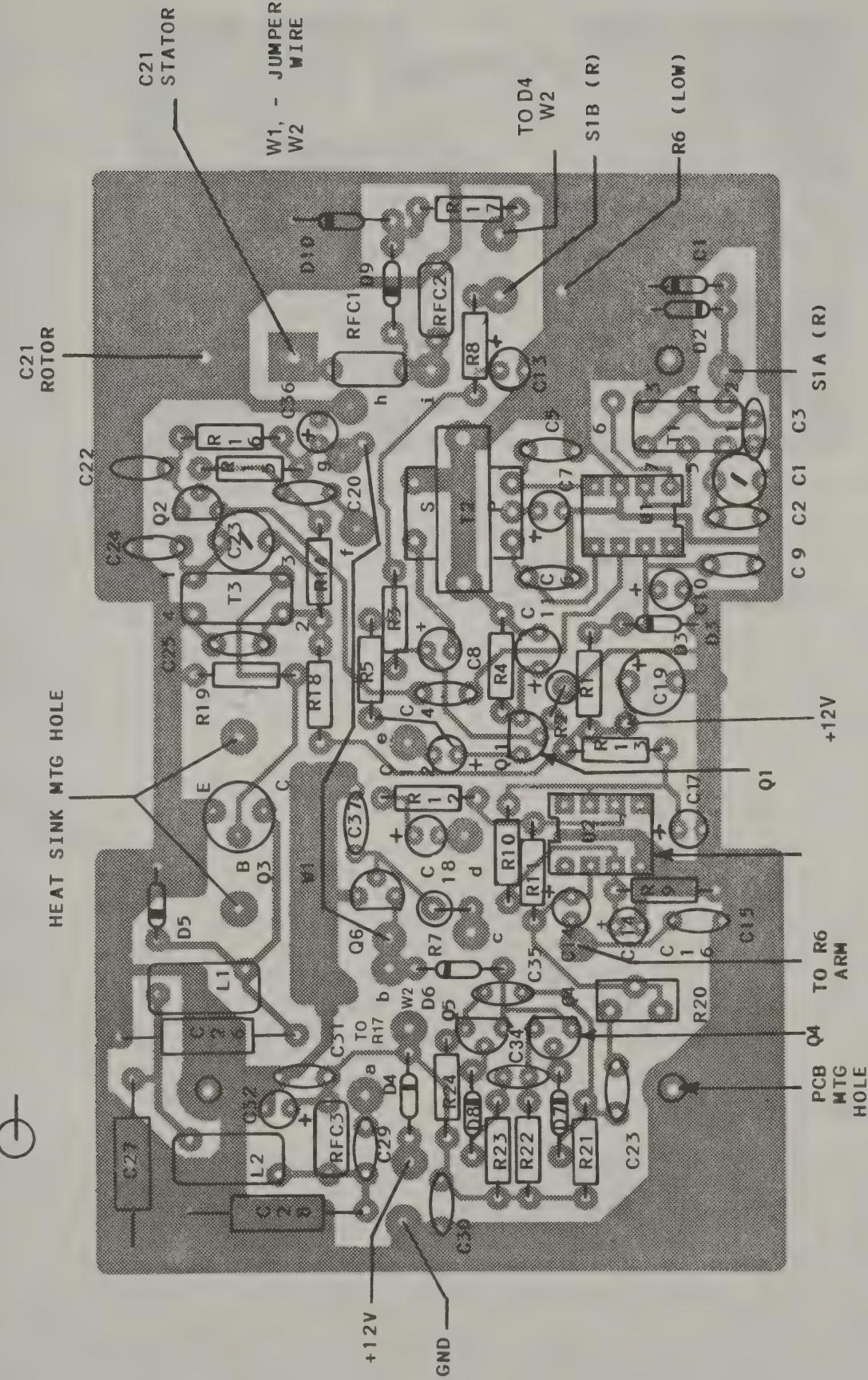


Fig. 6-14 -- Enlarged X-ray view of the PC board for the 40-meter transceiver as viewed from the component side of the board. W_1 may be omitted if both **b** and **g** are connected to J4. The black band indicated the cathode end of the diodes. The electrolytic capacitors should be installed in accordance with the polarity shown.

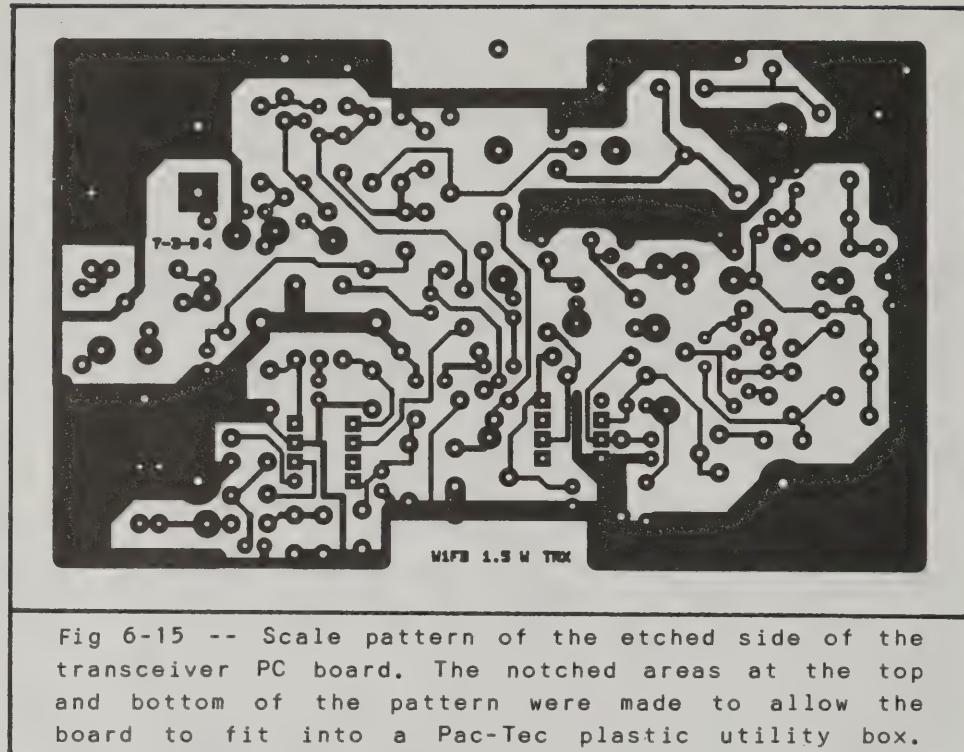


Fig 6-15 -- Scale pattern of the etched side of the transceiver PC board. The notched areas at the top and bottom of the pattern were made to allow the board to fit into a Pac-Tec plastic utility box.

The PC board in Fig 6-15 is available as a drilled and plated item from FAR Circuits, 18N640 Field Court, Dundee, IL 60118. I do not recommend that you attempt to build this project on the universal PC board described earlier in this book. Perf Board or "ugly construction" is not recommended either.

CW selectivity may be enhanced considerably if you use a two- or three-pole RC active audio filter at the output of the receiver (J1 of Fig 6-11). Fig 5-18 shows the circuit for a four-pole RC active CW filter that may be used with this transceiver.

Many of the components for the 40-meter transceiver are available by mail from Oak Hills Research, 20879 Madison Ave., Big Rapids, MI 49307 (KE8KL). For example, mini trimmers C1 and C23, the transistors, RF chokes, diodes and numerous other components are available from that source. A catalog is available in return for an s.a.s.e. with 50 cents U.S. postage. QRP CW transmitter kits are also available from KE8KL.

Double Sideband Suppressed Carrier Gear

We can build simple DSB (double sideband) equipment for experimental low-power applications. I certainly do not recommend this type of gear for day-to-day fixed-station use. Spectrum conservation rules out the use of any double-sideband equipment in our HF bands. Good carrier suppression (50 dB or greater) is also important in terms of spectrum conservation. However, QRP DSSC transmitters may be useful to the hiker, camper or user of emergency equipment. The relative simplicity of DSSC gear, along with the low cost of the package, appeals to many builders. Carrier suppression may be as great as 40 dB, but more typically it is on the order of 30 dB. Equipment that produces less than 10 watts of RF output power is seldom offensive when operated in the DSSC mode. QRM from the unsuppressed sideband is minimal unless you live very close to another ham who chooses to operate near your

frequency. It is important to be aware that two DSSC transmitters and two DC receivers in a single communications circuit are unsatisfactory. Either one is suitable, however, when used with a station that is equipped for SSB transmissions or reception. The lack of compatibility between two DSSC transmitters and two DC receivers results from the transmitter producing both USB and LSB energy while the DC receiver responds to or copies both sidebands at the same time.

The advantage of DSSC transmission is that the signal is developed at the chosen operating frequency rather than being heterodyned to a selected frequency. Also, there is no crystal or mechanical filter used. The lack of a filter is part of the reason why the carrier suppression in a DSSC transmitter is not as effective as it is in an SSB transmitter. In other words, the filter gets rid of some of the carrier energy along with the balanced modulator.

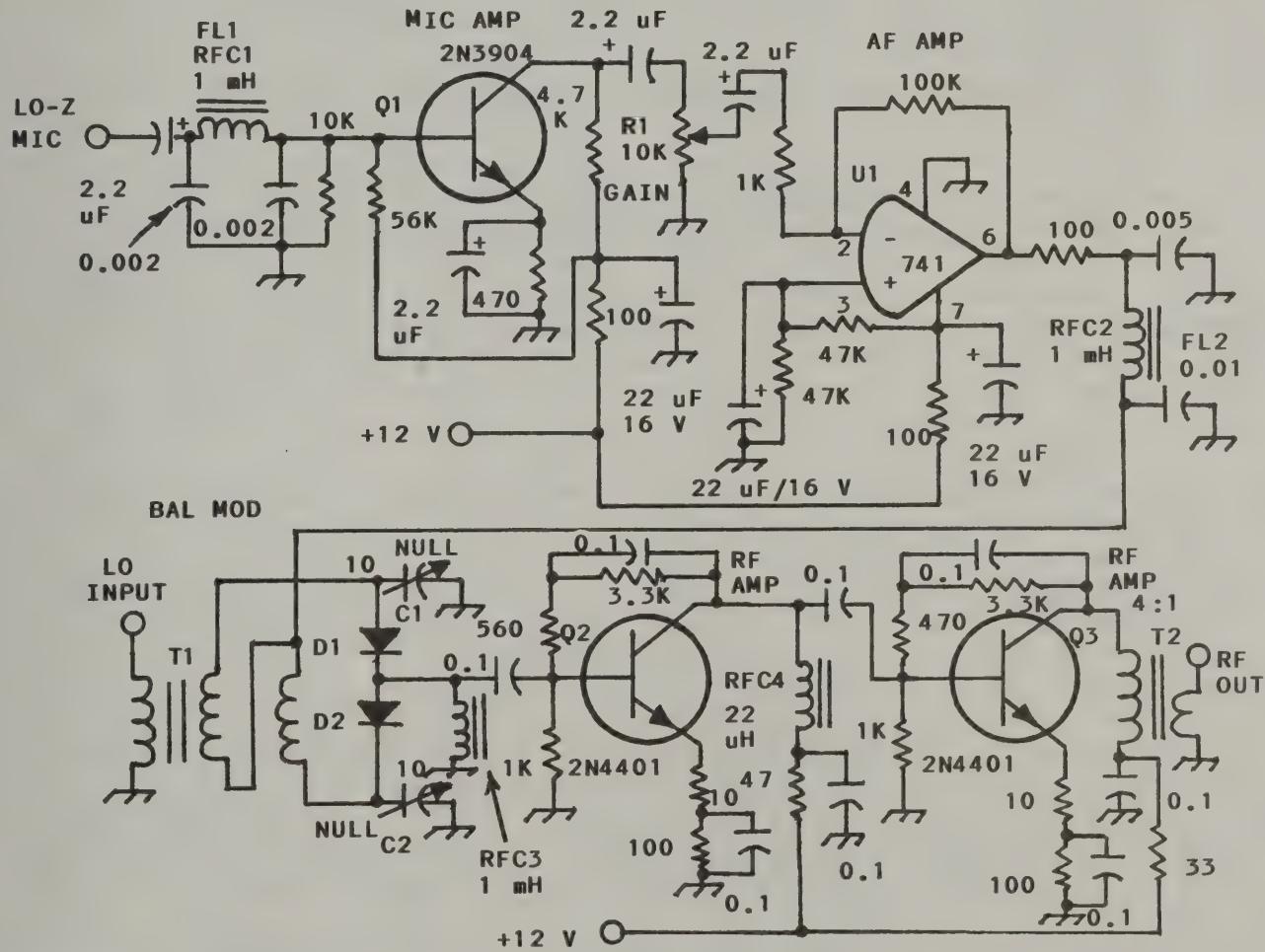


Fig 6-16 -- Circuit for an experimental DSSC exciter. Capacitors are disc ceramic. Polarized units are tantalum or electrolytic. Resistors are 1/4-W carbon film or composition. C1 and C2 are small ceramic trimmers. D1 and D2 are HP-2800 hot-carrier diodes or 1N914s matched for fwd resistance. R1 is an audio taper carbon control. T1 has 12 trifilar turns of no. 30 enam. wire on an FT-37-43 toroid. T2 has 12 pri turns of no. 30 enam. on an FT-37-43 ferrite toroid. Sec has 6 turns of no. 30 enam. wire.

Fig 6-16 shows a circuit I have used as a low-level DSSC generator. FL1 prevents RF energy from entering the audio circuit via the mic cord. It does not impair the frequency response of the audio system in the voice communication range. Op amp U1 delivers audio energy to the balanced modulator, D1 and D2. FL2 prevents RF energy in T1 from flowing into along the audio line. D1 and D2 should be closely matched in order to ensure good carrier suppression. The T1, D1 and D2 circuit layout needs to be as symmetrical as the art permits to aid carrier suppression. C1 and C2 compensate for circuit imbalance. Only one trimmer may be required. Adjust C1 and C2 for maximum carrier suppression, starting with the trimmers at minimum capacitance. You may eliminate the trimmer that has the least or no effect. Q2 and Q3 are linear broadband amplifiers. With the feedback networks shown they have an input impedance of 50 ohms and a collector impedance of 200 ohms. Q1 and Q2 are capable of some 15 dB of gain each. Output from T2 is at 50 ohms. Power output from Q3 is approximately 50 mW.

The Fig 6-16 circuit may be used as shown from 1.8 to 30 MHz. Carrier suppression becomes more critical (layout and adjustments) as the operating frequency is increased. The LO input power for T1 is +7 dBm (5 to 10 mW is okay). You may use a VFO or crystal oscillator for the LO system. A third-overtone oscillator, followed by a straight-through 2N4401 amplifier, is necessary for crystal control on 10, 15 or 17 meters.

You will need to add some linear power amplifiers after Q3 in order to obtain enough signal capability for communications. I suggest that you operate the Fig 6-16 DSSC generator into a class A 2N3553 or 2N3866. Examples of two-stage power amplifiers for DSSC 2- and 6-meter transmitters are presented in **Solid State Design for the Radio Amateur**, chapter 8. A harmonic filter is required at the output of the final amplifier of the overall system. All amplifiers used after Q3 of Fig 6-16 must be of the linear class. A suitable 5-watt output stage for the system is an MRF-476 or a 2SC1919.

I suggested earlier that a DSSC transmitter can be used as a portable or emergency rig. If you contemplate such a venture it is important that you keep in mind the dc current required by a group of linear amplifiers. The Fig 6-16 circuit alone draws approximately 40 mA at +12 volts. Two additional power stages (depending upon the chosen output power) can increase the current drain substantially. This should present no problem when operating from a boat or automobile battery. It can be crucial when working with dry cells or a NiCd battery pack.

Carrier suppression can be maximized by using careful circuit-board layout. The local oscillator should be contained in its own shielded compartment or box to prevent stray RF energy from reaching the output side of the balanced modulator. It is wise to enclose the balanced modulator in a small shield compartment to aid the isolation. I was able to obtain 35 dB of carrier suppression with the circuit in Fig 6-16.

Generating SSB Energy

A single-sideband generator is not vastly different than the DSSC generator. The main difference is that either the upper or lower sideband is eliminated to cut the bandwidth of the signal in half, so to speak. A mechanical or crystal filter is used after the balanced modulator, and the carrier generator (crystal osc.) produces two frequencies -- one for lower and one for upper sideband. These carriers are on frequencies that fall approximately 20 dB down on either side of the filter

response curve. A typical SSB filter has a 3-dB bandwidth of 2.4 kHz. For a 9.0-MHz SSB filter we would generate carriers at 8998.5 and 9001.5 kHz to facilitate upper and lower sideband operation. Amplified audio energy from the mic is applied to the balanced modulator as it is in Fig 6-16. The SSB suppressed-carrier signal is amplified, fed to a mixer and heterodyned by way of a local oscillator to the desired operating frequency. RF amplifiers follow this mixer to build the signal to the required output level.

Individual circuits for SSB generators are described earlier in this book. A number of practical composite SSB-transmitter circuits are published in **Solid State Design for the Radio Amateur**. Additional information may be found in **The ARRL Handbook**.

Measuring RF Power and SWR

An SWR bridge circuit appeared in the winter 1989/1990 edition of **SPRAT**, the ham journal of the G-QRP Club. It was designed by D. Stockton, GM4ZNX. I experimented with his circuit and found the performance to be excellent. The circuit is simple and bridge balance is easily achieved without the usual balancing trimmers found in the classic Bruene bridge. The circuit yields good accuracy, has a low insertion loss and is not frequency sensitive. It can be tailored for power levels from QRP to the legal amateur limit by adjusting the turns ratio of both of the transformers equally. Parts kits for this bridge are available from Kanga Products, 3 Limes Rd., Folkstone, Kent CT19 4AU, England. **SPRAT** is available to club members. Membership information is available from Rev. George Dobbs, G3RJV, the editor of **SPRAT**. Be sure to include sufficient IRCs for air mail when you request information.

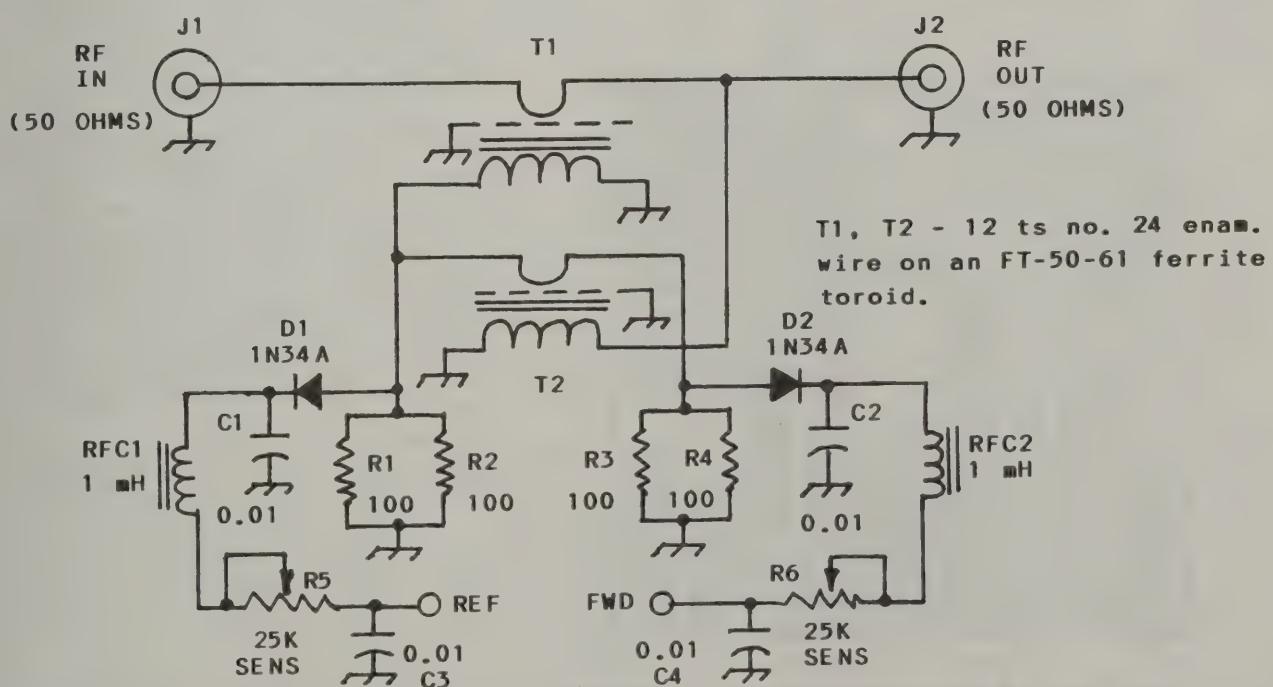
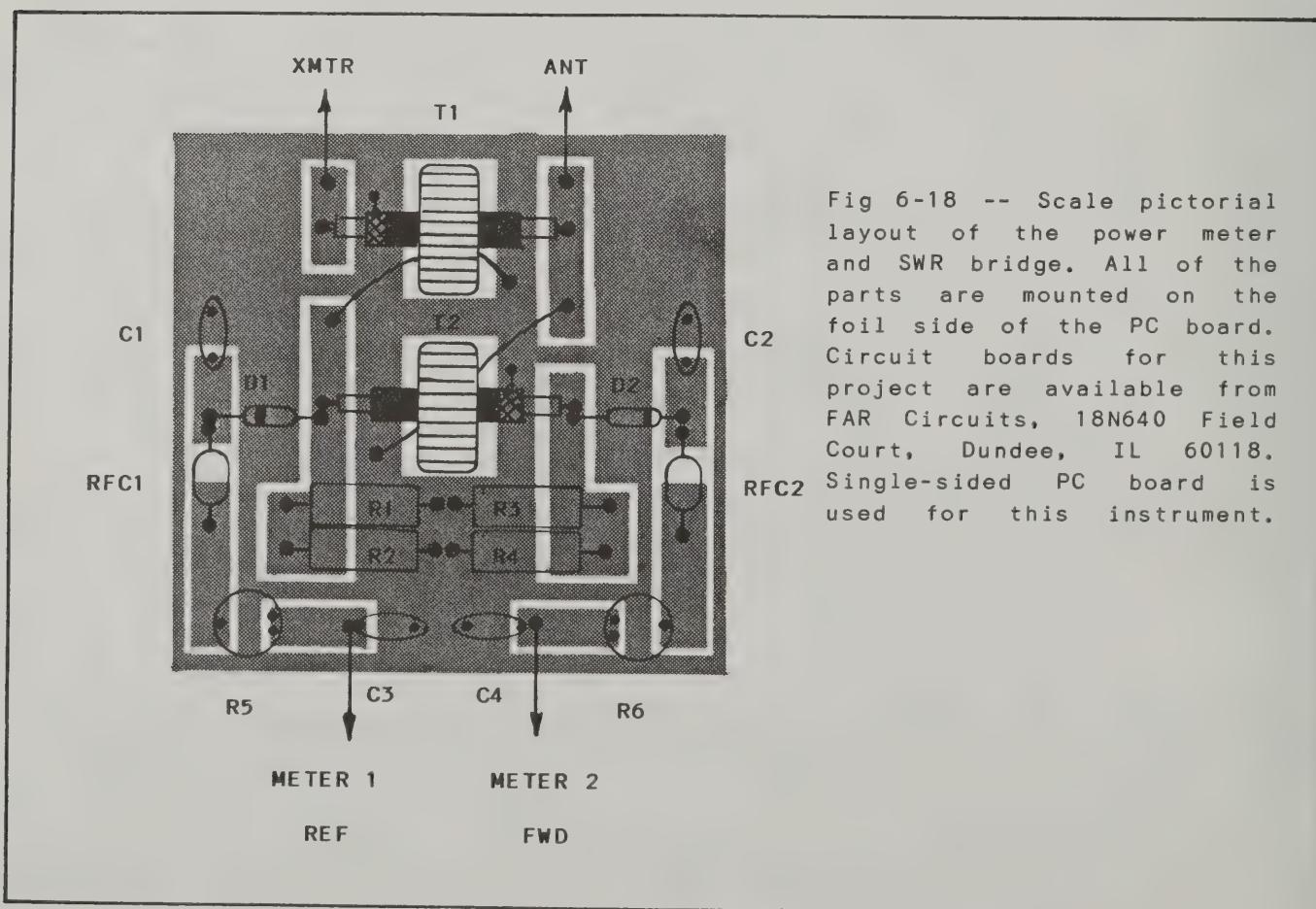


Fig 6-17 -- Schematic diagram of the RF power meter and SWR bridge. Capacitors are disc ceramic. R1-R4, incl., are 1-W carbon composition (noninductive), or equivalent combinations of 1/2-W carbon composition resistors. R5 and R6 are PC-mount carbon composition trimmers. T1, T2 specifications above are for a 1- to 150-W power range. Increase to 20 turns for 1 kW.

You may use silicon diodes, such as 1N914s, for D1 and D2 in Fig 6-17 if you build the bridge for power levels above 100 W. It is preferable to use germanium diodes for QRP power measurements. This is because they conduct at lower voltage levels than is characteristic of silicon diodes. This effect can result in no meter reading when the reflected power is low, which suggests the SWR is 1 when it may be 1.5:1 or greater.

The FWD and REF terminals of the Fig 6-17 circuit are connected to 100- μ A meters. One meter indicates forward power and the other one reads reflected power. A single meter may be used with a FWD/REF SPDT switch may be used in place of two meters if you want to reduce the cost of the instrument. There is no reason why you can't take advantage of the availability and low cost of the many surplus 200- μ A edgewise meters that are listed in surplus parts catalogs. The reduced meter sensitivity will decrease the sensitivity of the instrument at the low-power end of its range. I am able to obtain a full scale meter response at 3/4 watt with the component values listed in Fig 6-17 while using a 100- μ A meter.

The T1 and T2 primary windings consist of a piece of RG-58 coaxial cable that passes through the centers of the toroid cores. The shield braid of each short run of coax is grounded at one end to form a Faraday screen. This prevents harmonic current from being picked up by T1 and T2. These unwanted currents can cause false meter indications when the instrument is not connected to a nonresonant resistive load. The dashed lines show the Faraday screens. Each coax section represents a single turn for the transformer primary windings.



The circuit board in Fig 6-18 need not be so large. The layout can be compressed to provide a much smaller instrument. Wide conductors are desirable in order to minimize stray inductance. I made my prototype PC board by removing the unwanted copper with a hobby motor and small cone-shaped abrasive bit. Strips of thin copper may be glued to a piece of Formica or similar insulating material in lieu of a standard PC board.

Instrument balance for this spinoff of a 4-port hybrid combiner should be no problem if you keep the signal leads short and adhere to a symmetrical layout. Testing is done by connecting a 50-ohm resistive load to J2. Apply RF power to J1 and observe the FWD reading. Adjust R6 for a full-scale meter reading. Next, reverse the cables that connect to J1 and J2. Apply power and set R5 for a full-scale indication. Reverse the cables two more times and observe the meter readings. You should see zero deflection on one of the meters while there is full-scale deflection on the other. This indicates correct circuit balance.

RF power calibration of the meters may be done by measuring the rms voltage across a 50-ohm resistive load (J2) and calculating the RF power. P-P readings taken with a scope may be used for this step if you do not have a VTVM and an RF probe. You will need to make a new meter scale in order to read RF power. Additional sensitivity controls (R5 and R6 circuit points) may be added if you want to build a multirange power meter. These controls may then be selected by means of a rotary switch.

Toroid cores for this project are available from Amidon Assoc., Inc., 12033 Otsego St., N. Hollywood, CA 91607. The remainder of the components, inclusive of the 1-W carbon composition resistors, is available from the many Newark Electronics outlets in the USA. A list of distributors is available from the central office at 4801 N. Ravenwood Ave., Chicago, IL 60640-4496. Check your phone book Yellow Pages for the Newark outlet in your area.

Chapter Summary

I have attempted to lay a foundation upon which you can build a base for further learning of transmitting principles and practice. Certainly, this chapter is far from being all-inclusive. I encourage you to study **The ARRL Handbook** and **Solid State Design for the Radio Amateur** in quest of additional knowledge that is associated with the topics treated in this volume. **The W1FB QRP Notebook** is still another source for plain-talk theory about receivers and transmitters.

The "learn by doing" philosophy will work for you if you're willing to take that soldering iron in hand and experiment. Start with a simple project if you lack experience. Work with that circuit until you have it perfected, then move on to something more complex and make it function as it should. You will gain confidence and knowledge as you travel this exciting road to adventure.

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Appendix

More on crystal ladder filters

Selected articles from *QST*

A Unified Approach to the Design of Crystal Ladder Filters

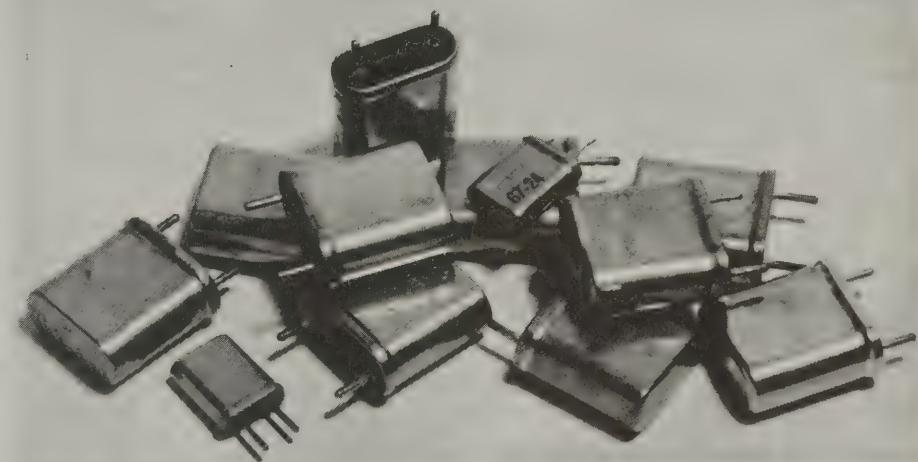
Have you turned away from a construction project because of high-cost crystal filters? Why not build your own?

By Wes Hayward, W7ZOI
7700 SW Danielle, Beaverton, OR 97005

The design of crystal ladder filters has been treated in the professional literature¹ and amateur journals.² However, the design methods are specific — they treat crystal filters as a special field with isolated methods. Actually, the same design methods for L-C bandpass filters may be applied directly to crystal filters. Such a unified design method should be more pleasing to the designer, and will allow more flexibility in the resulting filters.

The amateur's interest is primarily one of economics. Crystal ladder filters are easily designed and built with readily available components, yielding great savings for the builder. Television color-burst crystals are attractive for ladder filters. These are at 3.579 MHz in the U.S., while European crystals are at 4.433 MHz. The latter are popular with the many builders in the G-QRP-Club.

This paper addresses a number of goals. Simple methods are presented for the crystal evaluation and measurements needed for the design of filters. The accuracy is adequate for most amateur filter designs. Also, a simple set of design equations is given, which allows the measured data to be used in the design of filters. Tables are presented for Butterworth and



0.1-dB ripple Chebyshev filters.

Finally, some design subtleties are presented to aid in the construction and tuning of rather precise filters. These details may be ignored for many simple amateur-built filters, but should be of interest to the exacting designer. The results of the more exact methods are compared with the simplified ones.

All of the design may be done with a hand-held scientific calculator. Detailed analysis of filter frequency response may be done with sophisticated programmable calculators, such as those offered by Hewlett-Packard or Texas Instruments. All of the analysis reported by the writer was done with an HP-41CV calculator.

Some Filter Fundamentals

Fig. 1A shows a traditional LC-coupled resonator filter. This circuit uses two coupled, parallel-tuned circuits. Many more resonators (tuned circuits) may be used to obtain a steeper skirt response. Filter bandwidth and response shape are determined by the loading caused by the end terminations (the source and load resistances) and by the coupling between resonators.

There is no reason to restrict the filters to those using parallel resonators. Series-tuned circuits are just as viable. This type of L-C filter is shown in Fig. 1B, the exact duplicate of that using parallel resonators. Design of multielement filters of both types is covered in the literature.³

A detail that is not generally appreciated is the relative freedom available to the filter designer. For example, a 5-MHz L-C filter with a 100-kHz bandwidth could be designed with inductors of less than 1 μ H, with inductors greater than 10 μ H, or with anything in-between. Some values might be more practical, but this is not a fundamental restriction. Once an inductor is chosen, the rest of the filter components are determined. This applies to both filters of Fig. 1.

Another overlooked detail can be the termination of the filters. Any filter *must* be terminated at both ends in the resistance for which it was designed. The filters of Fig. 1 are doubly terminated with equal resistances at each end. This is common, but not mandatory. It is not proper, however, to design a filter for a given load at each end, such as 50 ohms, and then to expect the same response from that filter with other terminations.

Fig. 2A shows the equivalent circuit for a quartz crystal. This is a model — a circuit that shows the same response as a real crystal. The components of the model may not be practical, but this is of no significance for design work. For example, a 5-MHz crystal used in some filters built by the writer had a motional inductance of $L_m = 0.098$ and a motional capacitance of $C_m = 0.0103$ pF. The loss resistance was $R_s = 13.4$ ohms, and the parallel capacitance was $C_p = 5$ pF.

The parallel capacitance, C_p of Fig. 2A,

¹Notes appear at end of article.

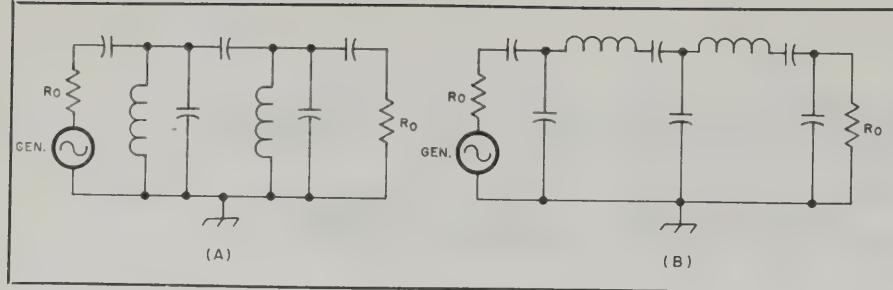


Fig. 1 — Double-tuned circuits with equal termination at each end. Parallel resonators are used at A; series-tuned circuits are used at B.

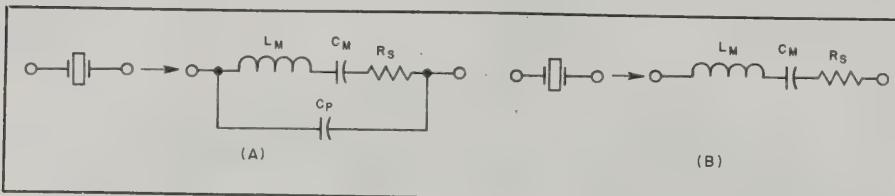


Fig. 2 — Equivalent circuit for a quartz crystal. L_m and C_m are "motional" components. Parallel capacitance, C_p , is included at A while it is ignored at B.

is rarely more than a few picofarads and may often be ignored for a design. This leaves the equivalent circuit of Fig. 2B. This is nothing more than the series-tuned circuit, exactly like that used in the L-C filter of Fig. 1B. Realizing this, the same methods may be used for the design of an L-C or crystal filter — one is no more complicated than the other.

Crystal Measurements

If the methods of LC-filter design are transferred to the design of crystal filters, it is mandatory that the vital inductance and capacitance values in the crystal, L_m and C_m , be known. It is not reasonable to substitute an arbitrary crystal of "proper" frequency into an existing design and expect it always to work. This will be illustrated later.

The crystal parameters are measured indirectly, but easily, with lab-quality instrumentation. A suitable measurement may also be done with equipment available to most amateur experimenters, and with a special test set constructed from ordinary components.

The nature of the measurements is understood with reference to the simplified crystal equivalent circuit of Fig. 2B. The crystal is placed between a source of known characteristic impedance and a detector, also of well-defined impedance. The generator is adjusted until a peak response is found. Then, the reactances of L_m and C_m cancel, leaving the result dominated by R_s . This is evaluated easily by replacing the crystal with a small-value variable resistor that is adjusted for the same response in the detector. The potentiometer is then measured with an ohmmeter, providing a value for R_s .

Next, the crystal is reinserted in the signal path and the generator is tuned to both sides of center frequency. The two frequencies where the response is down by 3 dB are noted. The difference is the loaded bandwidth in the test circuit. This is used to calculate a loaded-Q value. But, this is directly related to L_m . Using the condition for resonance, C_m is then calculated. C_p may be measured, but is not vital to the design of the filters described in this paper. Instead, we have assumed that $C_p = 5 \text{ pF}$ for all examples.

Fig. 3 shows the test set that is used to perform the measurements. The first element is a signal generator. It should have an adjustable output level (up to about -10 dBm or more) and should have excellent stability and good bandspread. Remember that we may be measuring frequency differences of only 100 Hz or so. A suitable generator is described in Chapter 7 of *Solid State Design*.⁴ The output of the signal source is applied to a frequency counter and to the test set. The counter should have a 1-Hz resolution.

The generator output is attenuated with a 20-dB pad and then applied to the crystal. The high attenuation ensures that low power is delivered to the crystal and provides a 50-ohm termination for the crystal. Output from the crystal under test drives an amplifier with a 50-ohm input resistance. The signal is amplified by four gain stages and then applied to a diode detector, D1. The dc output drives a high-impedance voltmeter. The test set should be operated with output voltages of 2 or less to prevent overdrive of the amplifiers. Q4, the related components, and the detector may be eliminated, if desired. Then, the output from Q3 is routed to an

oscilloscope with a 50-ohm terminator.

Amplifier gain is switchable with S1. With S1 open, the net gain is somewhere around 40 dB. Closing S1 changes the emitter degeneration in Q2, causing the net gain to increase by 3 dB.

A batch of crystals may be evaluated easily for filter applications with this test set. A crystal is inserted and the generator is tuned for a peak response in the voltmeter. That response is carefully noted. The crystal is then removed and replaced with the potentiometer. This is adjusted to obtain the same voltmeter response as was obtained with the crystal. The "pot" is then removed from the circuit and measured, providing a value for R_s .

The crystal is now reinserted in the circuit, with S1 open. The generator is tuned again for a peak response. Both the series-resonant frequency, F_o , and the meter response are carefully noted. S1 is then closed to produce an increase in output. The generator is tuned to the two sides (above and below F_o) until the meter reads the same as it did earlier. The two frequencies are noted and the difference is recorded as Δf , a parameter used for later calculations. Note that there is no need for amplitude calibration anywhere in this system.

This procedure is repeated for a reasonable sampling of the crystals on hand. The work that the writer has done would suggest that the values for F_o , R_s and Δf may be averaged for later calculation as long as the spread is not excessive. One batch of 20 surplus TV color-burst crystals showed an average R_s value of 20.78 ohms, with a standard deviation of 7.2 ohms. The average series-resonant frequency was 3577.257, with a standard deviation of only 67 Hz.

The designs which follow are based on having all crystals in a filter at the same frequency. Hence, frequency matching is required. A rule of thumb is that the deviations should be less than about 30% of the bandwidth of the filter. The center frequency measured in the test set will be the series-resonant value. This is not necessarily the value that would come from an oscillator. A simple oscillator is shown in Fig. 4. It may be used for matching the crystals. This circuit operates at a frequency slightly higher than the series-resonance. It is still suitable for frequency matching. It will serve also for BFO applications. The operating frequency may be increased further by insertion of a variable capacitor in series with the crystal.

Simplified Filter Design

Now that data is available on existing crystals, filter design may commence. See Fig. 5. A few approximate equations may be used. They require additional data and normalized filter parameters. These are presented in Tables 1 and 2, respectively,

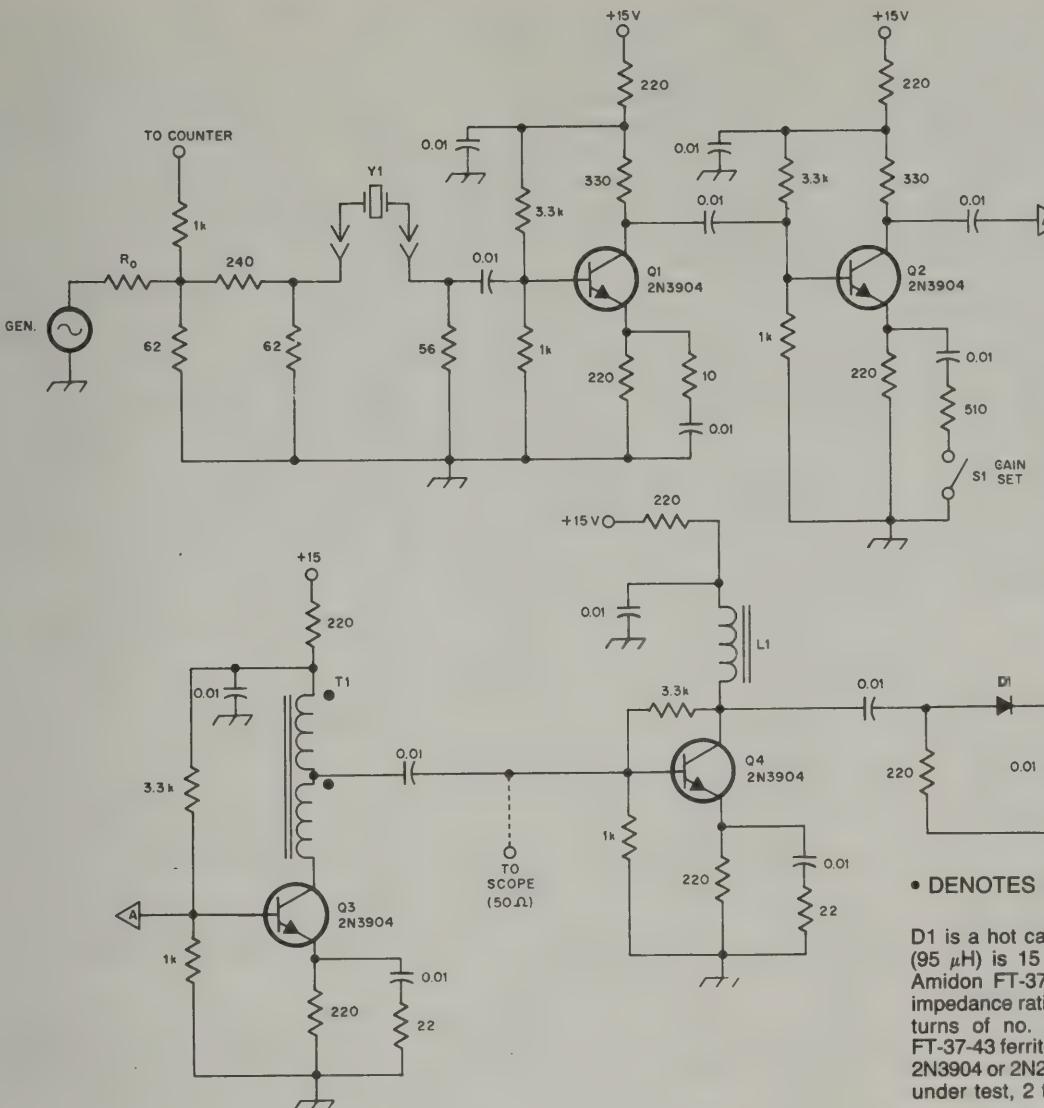


Fig. 3 — A simple test set for the evaluation of crystals to be used in filters. Construction is not critical.

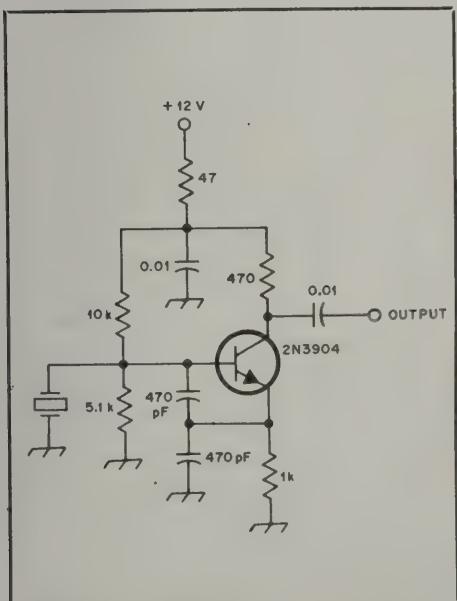


Fig. 4 — A simple crystal oscillator that may be used for crystal frequency matching. The 470-pF capacitors may be ceramic, mica or polystyrene. This circuit will function with crystals from 1.8 MHz to over 10 MHz.

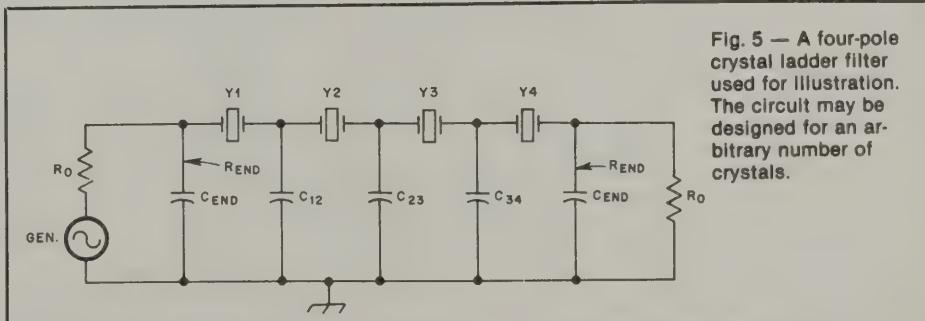


Table 1
Normalized k and q Values for a Butterworth Response Without Predistortion

<i>N</i>	<i>q</i>	<i>k</i> ₁₂	<i>k</i> ₂₃	<i>k</i> ₃₄	<i>k</i> ₄₅
2	1.414	0.7071			
3	1	0.7071	0.7071		
4	0.7654	0.8409	0.4512	0.8409	
5	0.6180	1	0.5559	0.5559	1

Table 2
Normalized k and q Values for a 0.1-dB-Ripple Chebyshev Filter Without Predistortion

<i>N</i>	<i>q</i>	<i>k</i> ₁₂	<i>k</i> ₂₃	<i>k</i> ₃₄	<i>k</i> ₄₅
2	1.6382	0.7106			
3	1.4328	0.6618	0.6618		
4	1.3451	0.685	0.5421	0.685	
5	1.3013	0.7028	0.5355	0.5355	0.7028

for the Butterworth and the 0.1-dB ripple Chebyshev filters. More exacting designs may be done with the exhaustive tables presented by Zverev.⁵ His work considers the effects of response-shape distortion, a consequence of loss in filter elements. The Zverev tables are termed "predistorted."

The motional crystal components, L_m and C_m , have been factored into the equations and need not be calculated. Equations are given for them, though. An equation is also given for the unloaded crystal Q (Q_u). This parameter is not needed directly for filter design, but should be evaluated, nonetheless. The Q_u value should exceed the filter Q by a factor of 10 or more to allow simple filters to be built. Filter Q is defined by F_{center} + bandwidth, where both are in hertz. Some surplus crystals may have a Q_u value that is too low for filter applications. The parameters are defined with respect to Fig. 5 as:

Δf	= bandwidth measured in test fixture (Hz)
B	= filter bandwidth in Hz
R_o	= end termination to be used (must be greater than R_{end})
R_{end}	= end resistance required to terminate the filter without matching capacitors
C_{end}	= matching end capacitor (pF)
C_m	= crystal motional capacitance (F)
L_m	= crystal motional inductance (H)
F_o	= crystal center frequency (MHz)
R_s	= crystal series-loss resistance as measured in test set
C_{jk}	= coupling capacitor (pF)

C_p	= crystal parallel capacitance (assumed to be 5 pF in all equations)
k_{jk}	= normalized coupling coefficient, given in Tables 1 and 2
q	= normalized end-section Q , given in Tables 1 and 2
N	= number of crystals to be used in the filter

The simplified design equations are

$$C_{jk} = 1326 \left[\frac{\Delta f}{B k_{jk} F_o} \right] - 10 \text{ (pF)} \quad (\text{Eq. 1})$$

$$R_{end} = \left[\frac{120 B}{q \Delta f} \right] - R_s \text{ (ohms)} \quad (\text{Eq. 2})$$

$$C_{end} = \left[\frac{1.59 \times 10^5}{R_o F_o} \right] \times \sqrt{\frac{R_o}{R_{end}}} - 1 - 5 \text{ (pF)} \quad (\text{Eq. 3})$$

Additional equations not mandatory for simple designs are

$$Q_u = \frac{1.2 \times 10^8 F}{\Delta f R_s} \quad (\text{Eq. 4})$$

$$C_m = 1.326 \times 10^{-15} \left[\frac{\Delta f}{F_o^2} \right] \text{ (farad)} \quad (\text{Eq. 5})$$

$$L_m = \frac{19.1}{\Delta f} \text{ (henrys)} \quad (\text{Eq. 6})$$

The design process will be illustrated with an example. Note that the equations use the units given in the list of parameters. Assume that a small group of crystals is frequency matched and found to have the average parameters $f = 294$ Hz, $\Delta F_o = 3.577$ MHz and $R_s = 23$ ohms. This data is typical of inexpensive TV color-burst crystals. These crystals will be used to design a 3-pole Butterworth crystal filter with a 250-Hz bandwidth. We eventually would like to terminate the filter in 50 ohms, but will not pick a termination, R_o , just yet.

The normalized coupling and loading values are found in Table 1 with $N = 3$. We see that $k_{12} = k_{23}$. Hence the coupling capacitors (Fig. 5) will be equal. The capacitor value is evaluated with Eq. 1.

$$C_{12} = C_{23} = 1326 \times \left[\frac{294}{250 \times 0.7071 \times 3.577} \right] - 10 = 606.5 \text{ pF} \quad (\text{Eq. 7})$$

The end resistance needed to terminate the filter is given by Eq. 2.

$$R_{end} = \left[\frac{120 \times 250}{1 \times 294} \right] - 23 = 79 \text{ ohms} \quad (\text{Eq. 8})$$

A value for R_o may now be picked. It may be any resistance greater than 79 ohms. A value of 200 ohms is chosen, and an end-

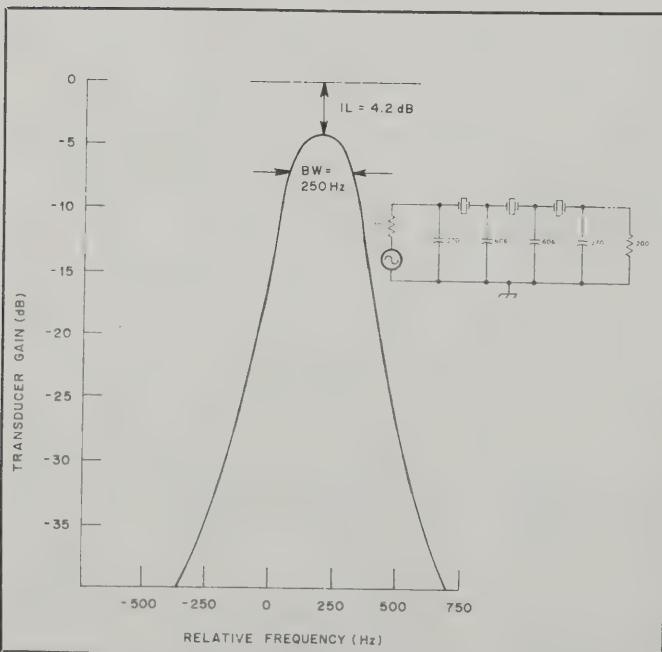


Fig. 6 — Circuit and calculated response for a 3-pole crystal filter at 3.577 MHz with a 250-Hz bandwidth. The design is based on measurements on surplus TV color-burst crystals.

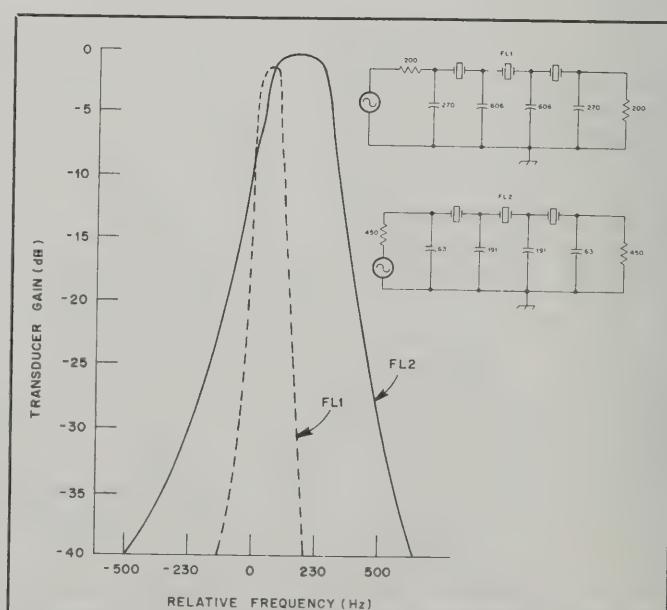


Fig. 7 — The response curves show the effect of using the wrong crystals in a design. The narrow response uses the circuit of Fig. 6 (FL1) with high-quality crystals. The wider response uses the circuit shown (FL2), designed on the basis of measurements on the crystals. Crystal data: $F_o = 3.579$ MHz, $R_s = 6.2$ and $Q_u = 721,000$.

matching capacitor is calculated from Eq. 3.

$$C_{end} = \left[\frac{1.59 \times 10^5}{200 \times 3.577} \right] \times \sqrt{\frac{200}{79}} - 1 - 5 = 270 \text{ pF} \quad (\text{Eq. 9})$$

The final circuit is shown in Fig. 6. The calculated response is also shown. These calculations were done using the Ladder Method,⁶ and accounted for the 5-pF C_p value.

The 200-ohm resistance levels may be transformed to 50 ohms with ferrite transformers. This may not be needed in a circuit application, but is certainly useful for measurements.

Note that the filter has an insertion loss of 4.2 dB. This results from the crystal loss with a relatively low crystal Q ($Q_u = 63,000$), evaluated with Eq. 4. The bandwidth is very close to the desired 250-Hz value, but the peak shape is much more rounded than would be expected from a Butterworth filter. This is also a result of crystal loss.

Consider now the effect of using the circuit of Fig. 6 with other crystals. A group of high-quality 3.579-MHz crystals were measured in the test set. The results were very different than those found with the surplus crystals. The Δf was 96 Hz, and R_s was only 6.2 ohms. The calculated Q_u was 721,000 — over 10 times that of the surplus crystals. L_m was also much different.

The filter of Fig. 6, designed for the low-Q surplus crystals, was evaluated with the parameters of the high-Q crystals. The

result is shown in the narrow response of Fig. 7. The bandwidth is much narrower than the desired 250-Hz value, and would be nearly useless in a cw receiver, owing to excessive ringing.

A filter was then designed around the measured crystal parameters. This response is also shown in Fig. 7, as is the circuit, FL2. This filter has a low insertion loss of only 0.4 dB and a shape like that expected of a Butterworth design. A comparison of Fig. 6 and Fig. 7 illustrates the effects of shape distortion.

The data in Figs. 6 and 7 are calculated. An obvious question regarding any experimental pursuit is how well do calculated curves compare with measured data? This comparison is presented in Fig. 8 for a 5-MHz, 250-Hz bandwidth filter. The three-pole circuit is also shown in the figure. The comparison between calculation and measurement is very good. The offset in center frequency is of no significance — it resulted from using 4.999 MHz for F_0 during the calculation, rather than a more accurate value. The measurements were done with the writer's receiver synthesizer as a signal generator, followed by a step attenuator and then the filter. This was followed by a broadband amplifier and a 50-ohm-terminated oscilloscope. Similar results have been obtained with filters at 3.579 MHz.

The experimental curve of Fig. 8 is marked with a BFO frequency. This would be the proper frequency to provide a 700-Hz beat note and to ensure good suppression of the opposite sideband. This filter would be very practical in a

simple superhet receiver for cw application, especially if it were supplemented with an R-C active low-pass audio filter. This scheme has been used very successfully by the writer in a portable Field Day transceiver.⁷

A simple three-pole filter is also practical for some ssb applications. Such a filter at 5 MHz is shown in Fig. 9. Measured and calculated frequency-response curves are also shown. Note that the filter shape is lacking in symmetry. Indeed, this is an illustration of why this type of circuit is termed a "lower sideband ladder." This also justifies the position of the BFO in Fig. 8. In spite of the poor attenuation slope on the low-frequency side, the filter of Fig. 9 would be practical for a simple ssb exciter. This passband ripple may be eliminated by tuning, a detail shown in the figure and covered in the following section.

Filter Tuning

The method presented so far has assumed that all crystals are exactly at the same frequency. This simplification is sufficient for many applications, especially with narrow-bandwidth filters. It is not adequate for critical designs. Additional tuning is required if we want to design filters with an exactly predicted bandwidth, achieve a desired shape more accurately, or build wide-bandwidth filters with more than 3 or 4 crystals.

Modern filter theory has been used to calculate coupling and end-matching capacitors. The detail that we have ignored is that the individual crystals have

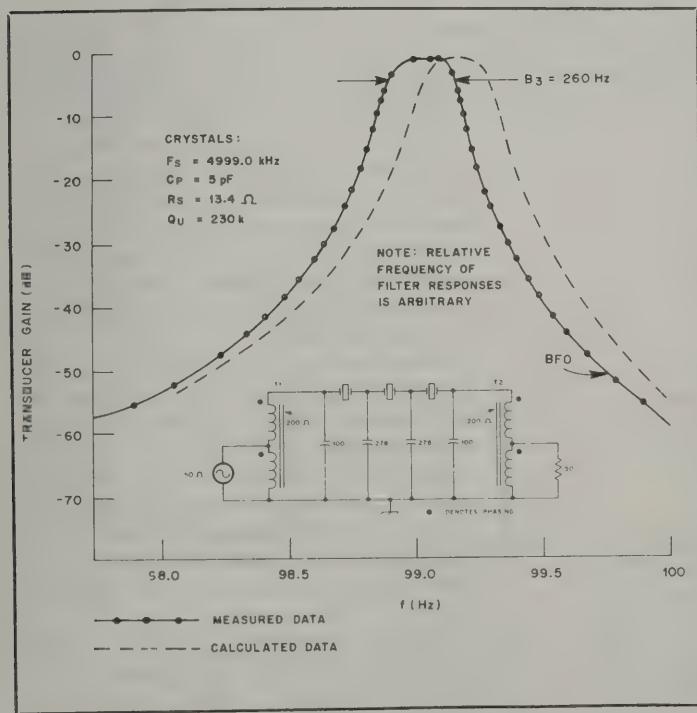


Fig. 8 — Comparison of calculated and measured results on a 5-MHz cw filter. The crystals had $F_0 = 4.999$ MHz, $C_p = 5$ pF, $R_s = 13.4$ ohm and $Q_u = 230,000$. A proper BFO frequency is marked. See text for discussion.

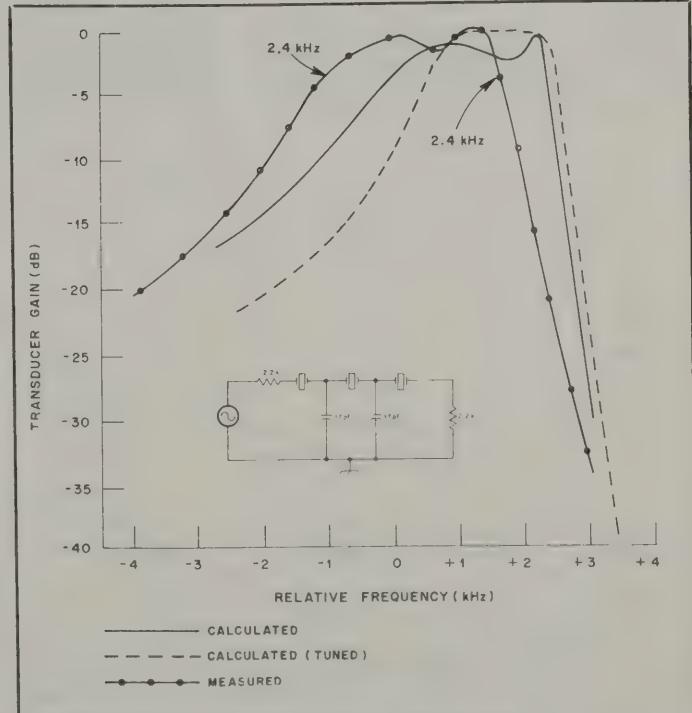


Fig. 9 — Measured and calculated (with and without tuning) results for a wider bandwidth filter for use in a simple ssb exciter. Crystals are the same as used for the filter of Fig. 8.

been detuned by the shunt capacitors. The detuning may be "fixed" through proper choice of crystal frequencies or with the insertion of additional capacitors in the circuit.

Fig. 10A shows one end of a crystal filter. The crystals have been replaced with the simplified equivalent circuit without loss resistance. All crystals are assumed to be at the same frequency, which implies that all L_m and C_m values are the same throughout the filter.

Filter theory states that each loop in this circuit should be resonant at the same frequency when that loop is considered alone. Adjacent loops are open-circuited during this evaluation. Consider the interior loop shown in Fig. 10B. The crystal resonance is determined by L_m and C_m . However, the resonant frequency of this loop is determined by L_m and the series equivalent of the three capacitors, C_{12} , C_{23} and C_m . The loop will resonate slightly higher than the crystal frequency. The actual frequency is easily calculated with standard formulas and a hand calculator.

Analysis of end-section resonance is slightly more complicated. The end resistance and the related parallel capacitance must be resolved into a series equivalent, shown in Fig. 10C. The series capacitance is given by

$$C_s = \frac{\frac{1}{R_o^2} + \omega^2 C_{end}^2}{\omega^2 C_{end}} \quad (\text{Eq. 10})$$

where $W = 2\pi F_o \times 10^6$.

The resonant frequency of the end loop is then calculated from L_m and the series combination of C_s , C_m and C_{12} .

Consider an example, a 4-pole ssb filter. This circuit, shown in Fig. 11, was designed for a 0.1 dB Chebyshev response using the k and q values from Table 2 and

crystals with $F_o = 3.577$ MHz, $R_s = 23$ ohms and $Q_u = 63,000$. Assume for the present that the two 96-pF capacitors are shorted. The filter is symmetrical, so only two frequency calculations must be done — one for the ends and one for the interior loops. The end loops are resonant 1220 Hz above F_o , while the inner two loops resonate 1790 Hz above F_o .

Two methods may be used to tune this filter, to force all loops to resonate at the same frequency. One method requires that the F_o of the crystals in the end sections be increased 570 Hz over that of the inner loops ($570 = 1790 - 1220$). This difference is small compared with the filter bandwidth, so we would not expect a dramatic difference in response shape.

The other method places capacitors in series with crystals in those loops requiring a frequency increase — the end sections of Fig. 11. This is the circuit shown in the figure. A value of 96 pF was found to be necessary for the 570-Hz shift. The addition of series capacitors allows more design flexibility, especially when working with surplus crystals. On the other hand, it may be convenient to use stagger-tuned crystals if the batch on hand has the proper frequency spreads. It is, of course, possible to use a combination of the two methods.

The frequency response of the four-pole filter is shown in Fig. 11 for the cases with and without tuning. The center frequency is raised with tuning, and passband ripple is reduced to near the desired 0.1-dB level. Either filter would be practical for amateur applications.

The bandwidth of the ssb filter of Fig. 11 was 2.2 kHz. It was designed for a 2.5-kHz bandwidth. The difference results from simplifying assumptions used to derive the critical equations and using the k and q values from Table 2, where the effect of filter loss is ignored during

design. Improved accuracy is obtained with the Zverev tables to supply the k and q values. Suitable amateur filters may be designed by increasing the design bandwidth slightly over the desired one, while using the Table 1 or 2 data.

The effects of shape distortion are more dramatic with narrow-bandwidth cw filters, for the losses are higher with decreased bandwidth. The Butterworth data of Table 1 provide a good starting point for design. As losses increase, the shape of a narrow-bandwidth filter evolves toward one with a more rounded peak shape, approaching something like a Gaussian response. This rounded shape is desired for narrow-band application, for it offers an improved time-domain characteristic with less filter ringing. Chebyshev filters should not be used for cw applications.

The effects of filter tuning were examined experimentally in the filter of Fig. 12. This was a 250-Hz wide design using the 5-MHz crystals applied in other filter experiments. A 5-pole design was chosen. The measured responses with and without tuning are also shown with the circuit. Without tuning, the attenuation slope on the high-frequency side of the response was poor. Tuning the filter improved the shape and reduced the insertion loss. The shape was still not the Butterworth response predicted. This was traced to variations in the crystal frequencies that had not been taken into account during the design. This filter is destined for use in a multiband portable transceiver.

Conclusions and Applications

Home construction of crystal filters is very practical, especially for the experimentally inclined amateur with the usual amount of instrumentation. Laboratory-grade equipment is definitely not needed. It is important, however, that the filters be carefully designed, and that the designs be based on the crystals to be used. Measurements are performed easily in the home lab to obtain the needed crystal parameters. None of the filter circuits presented in this paper is suitable for exact duplication.

Filter tuning may or may not be required, depending on the filter to be built. An interesting filter that would never require tuning is a two-pole circuit. This results from symmetry. Any detuning would be the same in both crystals. Improved stopband attenuation may then be obtained with a cascade of several filters with isolating stages of gain between them.

Another interesting special case is the three-pole filter with all crystals at the same frequency. The terminating resistance, R_o , may be set equal to the calculated R_{end} . There will then be no shunt capacitors at the ends. This filter may be tuned by placing series capacitors

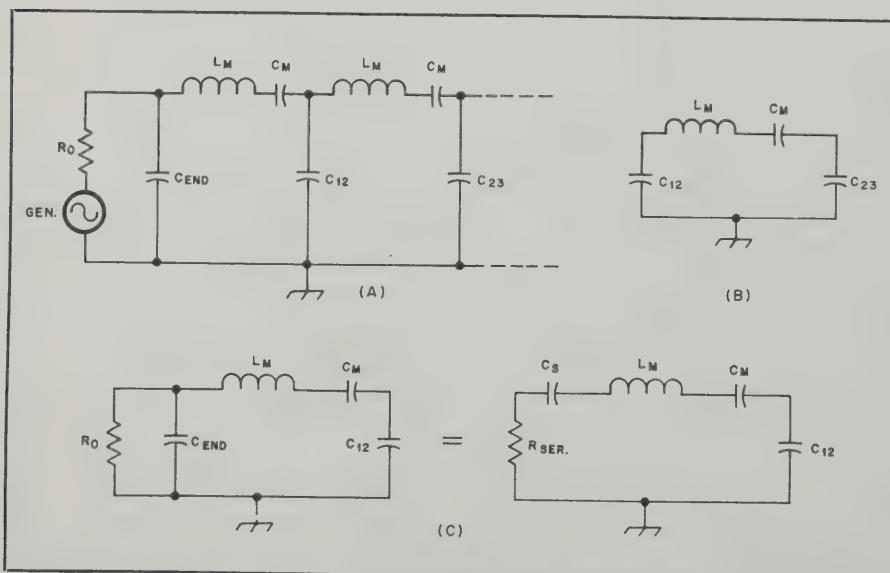


Fig. 10 — Partial filter circuits used in evaluation of tuning of individual loops. See text for details.

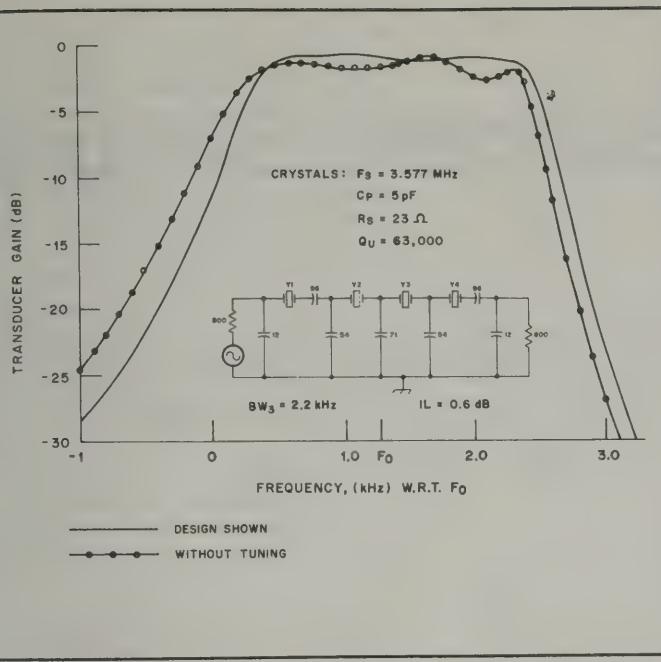


Fig. 11 — Calculated response for the 4-pole ssb filter shown. Surplus color-burst crystals were used for the design. The two response curves show the effects of filter tuning.

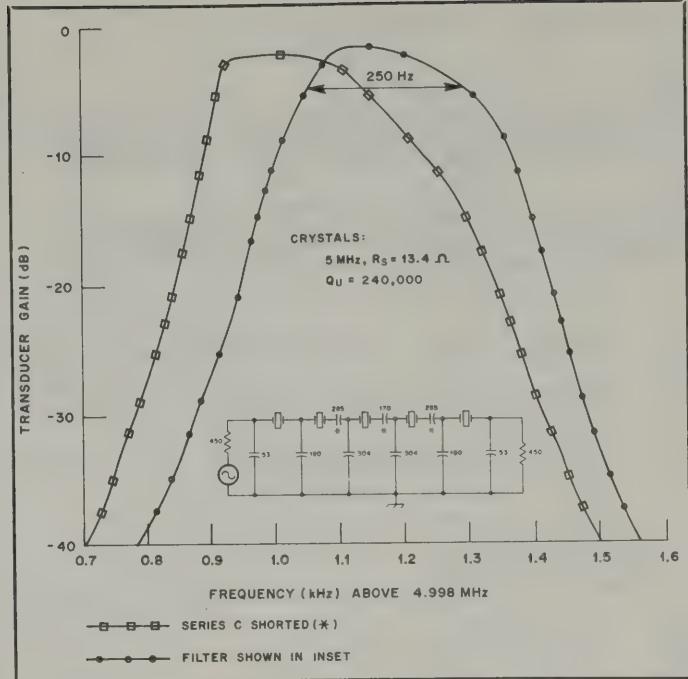


Fig. 12 — Measured response curves for a 5-pole, 5-MHz cw filter. The two curves show the effects of tuning a filter with series capacitors.

in the end loops, which have a value equal to the coupling capacitors. An example of this is the ssb filter shown earlier in Fig. 9. The tuning was realized with the addition of 17-pF capacitors in series with the two outside crystals. Variable capacitors may be used, of course, in any of the circuits shown. Instrumentation must then be built for alignment.

The 3.58-MHz TV color-burst crystals offer an attractive possibility for a simple cw receiver. A single local oscillator could be built at approximately 10.5 MHz. This will then allow both the 40- and 20-meter cw bands to be received with no band switching in the LO. This scheme is not

well suited to an ssb receiver, for harmonics of the BFO appear at 7.16 and 14.32 MHz. This harmonic relationship could also lead to spurious responses in a two-band cw transceiver.

Examination of the design procedure reveals some interesting subtleties. Careful choice of termination, R_o , could lead to filters requiring no tuning. Useful filters can be built from poorly matched crystals, although the design may get messy. Excessive tuning with series capacitors will increase the loss of the filter. Finally, careful choice of termination resistance will allow the construction of filters with a switched bandwidth. \blacksquare

Notes

¹A. I. Zverev, *Handbook of Filter Synthesis* (New York: John Wiley and Sons, 1967), Chapter 8.

²J. A. Hardcastle, "Ladder Crystal Filter Design," *QST*, Nov. 1980, pp. 20-23.

³W. Hayward, *Introduction to Radio Frequency Design* (Englewood Cliffs, NJ: Prentice-Hall, Inc., 1982). Both filter types are covered in detail in Chapter 3.

⁴W. Hayward and D. DeMaw, *Solid-State Design for the Radio Amateur* (Newington: ARRL, 1977), p. 171.

⁵See note 1, pp. 341-379.

⁶See note 3.

⁷See note 4, p. 214.

⁸The author has made arrangements to supply crystals at a variety of frequencies that have been characterized for filter applications. Interested readers should send an s.a.s.e. for a data sheet.

Designing and Building Simple Crystal Filters

A simple and inexpensive crystal filter that performs well makes receiver and transmitter projects much more fun. Build one yourself at a fraction of the cost of a commercial unit.

By Wes Hayward, W7ZOI

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Beaverton, OR 97005

I am encouraged by the large number of radio amateurs who want to build their own rigs. The ready availability of good-quality semiconductors helps in this pursuit. Other components are sometimes harder to find, at least at an affordable price. One example is the crystal filter—the heart of any superheterodyne receiver or transmitter.

Inexpensive crystals are readily available. They should be characterized and matched for frequency prior to use in a typical crystal filter. Methods for building the needed test equipment and performing the measurements have been presented before.¹ These methods are, unfortunately, somewhat complicated for the casual experimenter who may hesitate to construct special test equipment when just one filter is to be built. What experimenters really need is an empirical filter design method, one that lends itself to casual "tweaking." Such a method is described in this article.

The Cohn Filter

In the course of computer studies of both crystal and LC filters, I've noted that a circuit called the "Cohn," or "Min-loss" filter, lends itself to particularly simple designs.² This filter configuration derives its name from its originator, and differs from the more familiar Butterworth and Chebyshev circuits. The Butterworth bandpass filter is built for optimum flatness at the filter center. The Chebyshev design allows equal passband ripples, and is designed for the best stopband attenuation (steepest skirt response). The Cohn filter is a compromise: It is optimized to exhibit minimum insertion loss when built with practical resonators, while preserving a good shape factor. The Cohn filter, in LC form, is not new to the radio amateur.^{3,4} It is not limited to LC resonators, however. It works great with crystals!

The Cohn filter, crystal or otherwise, is a rather simple circuit. This becomes more apparent when we view the filter using coupled-resonator methods.⁵ All normalized coupling coefficients are equal. Moreover, the normalized end-section loaded-Q factor is the reciprocal of the coupling coefficient. The practical simplification becomes apparent if we examine the generalized crystal filter circuit shown in Fig 1. All capacitors in the circuit are of equal value! The shunt capacitors are coupling elements while the series capacitors in the filter end sections are included to properly tune the circuit.

Practical Cohn Crystal Filters

An empirical method that the amateur

may use for crystal filter design is described easily in a step-by-step procedure.

1) Obtain a collection of substantially identical crystals. The crystals are first matched in frequency. The same oscillator should be used to measure all crystal frequencies. The error (frequency difference) should be less than 10% of the *desired bandwidth* of the filter. For example, a filter with a 1-kHz bandwidth should use crystals matched to within 100 Hz or better.

2) Pick a capacitance value to be used in the filter. The capacitance (C) value determines the filter bandwidth. Larger C values yield narrower bandwidth and higher insertion loss.

3) Vary the end terminations to obtain a shape that is free of passband ripple while

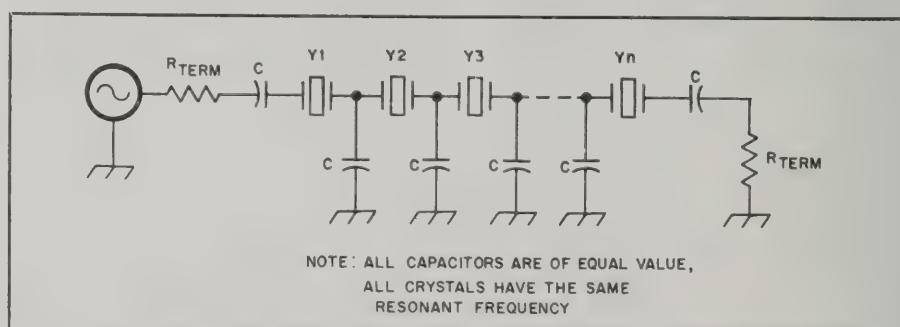


Fig 1—Generalized crystal filter suitable for empirical construction.

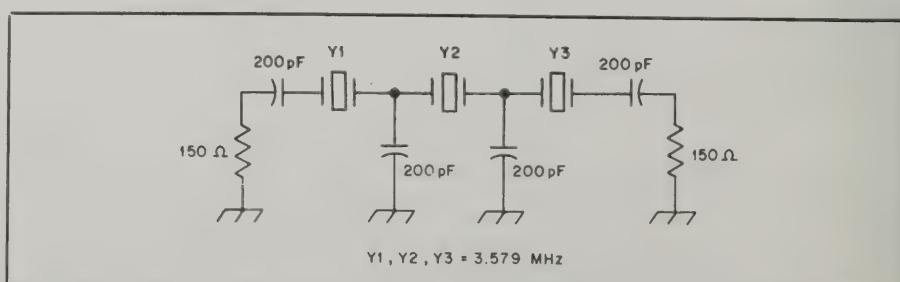


Fig 2—A simple CW filter using three crystals.

¹Notes appear at end of article.

providing sufficient stopband attenuation. This empiric procedure is illustrated in the following examples. I've cheated a bit—I used a personal computer to simulate the filter, and generate the data presented, but I've obtained similar results with filters I have built. The experimental results agree well with the computer models. All examples shown are based on a collection of crystals from my junk box. They are inexpensive 3.579-MHz TV color-burst crystals. The average motional inductance for these crystals is 117 mH, with a (rather poor) typical Q of 50,000. The parallel capacitance is about 4 pF.

A Three-Crystal Cohn Filter

A simple and practical filter for a beginner's first CW superheterodyne receiver is shown in Fig 2. Three crystals are used. The capacitors are 200-pF units, a standard value. Experimentation (done here with the computer) shows that a good filter shape is obtained with an end termination of 150 ohms. Fig 3 shows the frequency response of this filter. The -3 dB bandwidth is 403 Hz, and the insertion loss is 3.8 dB. The loss will be lower with better (higher Q) crystals. The impedance match is shown in the figure as a series of dots. This is the return loss normalized to the source impedance—150 ohms for the filter shown.

If different crystals are used, the same bandwidth can still be obtained, within limits. The coupling capacitors and end terminations will then be different, however. Insertion loss will also differ.

Decreasing the value of the capacitors increases the bandwidth. Some practical values are shown in Table 1, again the result of tweaking with the computer. This will provide some guidance in experimentation.

Fig 4 illustrates the effect of altering the terminating resistance. Fig 4A shows the result of 75-ohm terminations, lower than the desired 150-ohm value. The filter shows some passband ripple and a higher insertion loss. The effect of a 300-ohm termination is shown in Fig 4B, where the peak shape becomes more rounded, with degradation of skirt response. While the poorer frequency domain shape is generally less desirable, the filter with the higher termination has a significantly improved group delay; this filter would be preferred for high-speed data applications.

A Six-Crystal Cohn Filter

The three-pole filter mentioned above is practical. It does not, however, offer skirts that are as steep as we would like for many demanding applications. Improved skirt selectivity in a filter is obtained by using more crystals. The computer can be used to generate another table like that shown for the three-crystal filter. Alternatively, the results of Table 1 can be used as a starting point for experimentation. The

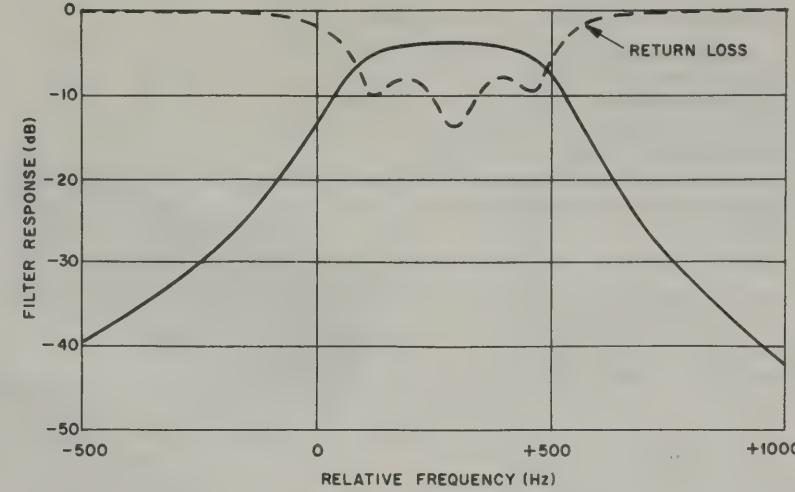
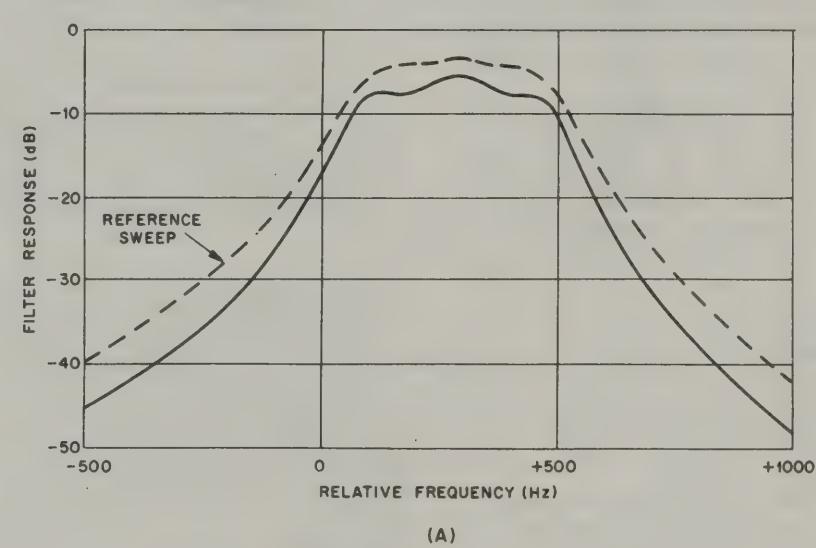
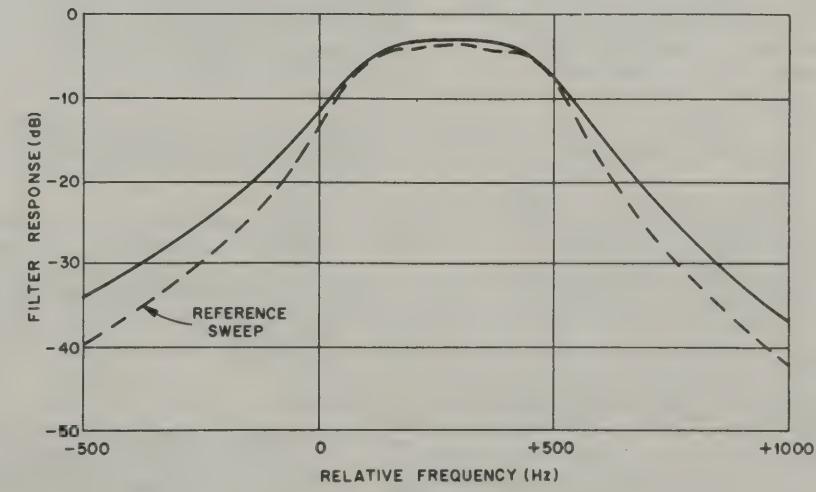


Fig 3—Frequency response of the filter of Fig 2. The dots show the input return loss, indicating the quality of the impedance match.



(A)



(B)

Fig 4—Frequency response of the filter of Fig 2 with changes in the end terminations. Curve A shows response with 150-ohm terminations; curve B shows the response using 300-ohm resistors. See the text for considerations of which is "better."

ARRL Lab Experiments with the Cohn Filter

ARRL Lab staff members were intrigued by the material on Cohn filters presented by Wes Hayward, W7ZOI. We built four CW filters and one SSB filter, following Wes's instructions. Tests confirmed the computer models developed by Wes. This was no surprise!

CW Filters

Four different batches of crystals were used for the CW filters. The crystal sources were identifiable, and the relative quality of each batch was determined. Four filters were constructed (Fig A). With the exception of the crystals used in each filter, the filters were identical. The filter schematic is shown in Fig B. The capacitors are 300-pF, 5%-tolerance silver-mica types. The 500-ohm terminations (variable resistors) at the ends of the filter were used to "trim" the filter for the best shape and response characteristics during testing. An HP-8540 spectrum analyzer was used to generate the filter response curves shown in photos C through G.

The units used in filter no. 1 are TV color-burst crystals (3.579545 MHz). They were purchased originally from Radio Shack (about \$1.60 each) for another project. There were only five of these crystals in the batch, so frequency matching (within 50 Hz) was not as close as with some of the other crystal batches.

The crystals used in filter no. 2 were selected from an assortment of ten 4.000-MHz microprocessor units purchased from JAN Crystals.¹ These crystals were frequency matched within 40 Hz. The crystals cost approximately \$3 each.

Filter no. 3 uses crystals selected on the basis of frequency matching from a large batch (over 30) of 4.000-MHz microprocessor crystals on hand in the ARRL Lab (matched within 30 Hz). These crystals can be characterized as "grab bag" quality, and similar units are available from various dealers at a cost of less than \$1 each.

We bought the crystals used in filter no. 4 from International Crystal Co.² They can be characterized as high-quality, moderate-cost units. Their guaranteed frequency tolerance is 0.001% of 4.000000 MHz, matching was within 6 Hz, and cost is approximately \$10 each.

SSB Filter

A four-crystal, 12-MHz SSB filter was built using 160-pF, 10%-

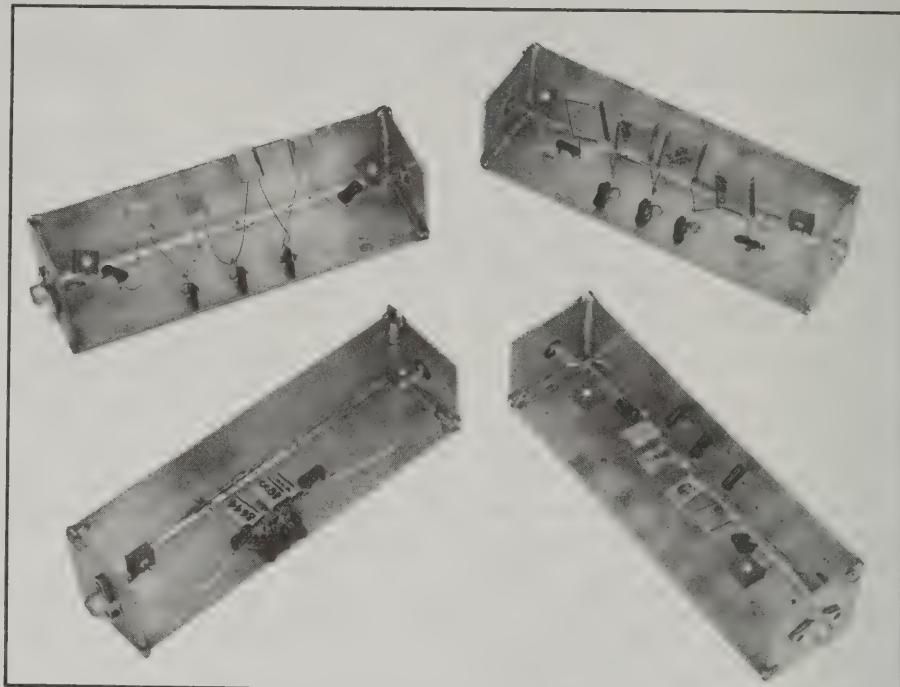


Fig A—Four CW crystal filters were built in this configuration. The PC-board mounting surfaces provide a ground plane. Capacitors are soldered directly to the ground plane, and the crystals are connected using the capacitors as standoffs. Phono jacks are used for input and output connectors. The only variables in the construction of the filters are the crystal characteristics and the length of the crystal leads. The SSB filter is not shown.

tolerance silver-mica capacitors. An 8:5 transformer is used for impedance matching. The crystals are microprocessor types purchased from Jameco Electronics³ at a cost of approximately \$1 each. Of 12 crystals purchased, only 10 were suitable for filter use. The filter response is shown in photo G.

Test Results

Photos C through F show the response curves of the four CW filters. Photo G shows the response curve for the 12-MHz SSB filter. Insertion loss is quantified only for

CW filter no. 4 because series resistors were used to adjust the terminating impedance of filter nos. 1 through 3. These resistors introduce losses. In practice, each filter would be coupled to its associated circuitry through matching transformers, not resistors.

Filter no. 1 exhibits an extremely sharp response, with a bandwidth of approximately 240 Hz at the -3 dB points; it may be too sharp for good CW copy. Changing the 300-pF capacitors in this filter to a lower value will broaden the response.

Filter no. 2 is not quite as sharp as filter no. 1, and exhibits a peak ripple effect. The response asymmetry can be corrected by trimming the filter

³Jameco Electronics, 1355 Shoreway Rd., Belmont, CA 94002, tel 415-592-8097.

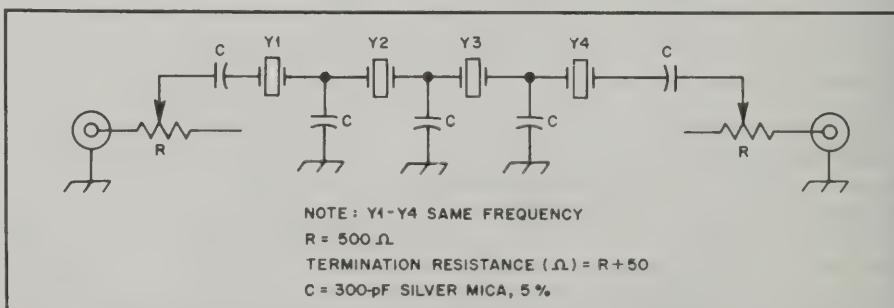


Fig B—Schematic diagram of the crystal filters. Capacitors are all of equal value. Terminating resistors are variable 500-ohm units. Crystals are all of equal nominal frequency with minor (up to 50-Hz) variation.

¹JAN Crystals, 2400 Crystal Dr., PO Box 06017, Fort Myers, FL 33906-6017, tel 813-936-2397.

²International Crystal Manufacturing Co., Inc., PO Box 26330, 701 W Sheridan, Oklahoma City, OK 73126-0330, tel 405-236-3741.

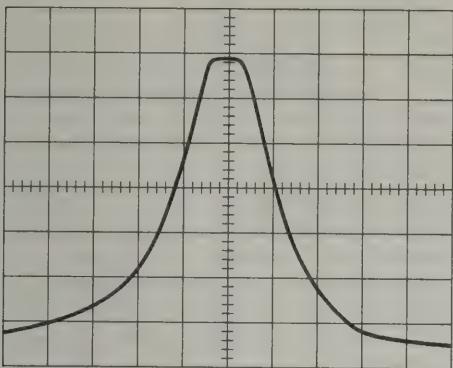


Fig C—Spectral photo showing the response of filter no. 1. Horizontal divisions are each 200 Hz; vertical divisions are each 10 dB. Sampling bandwidth is 100 Hz. The center frequency is 3.579 MHz.

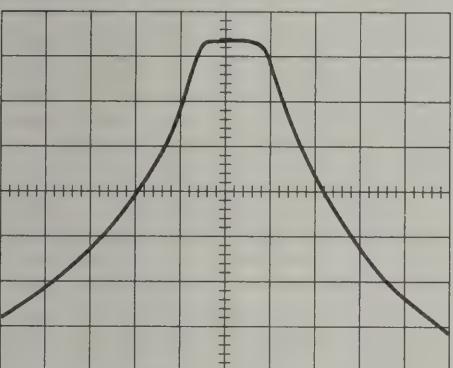


Fig D—Spectral photo showing the response of filter no. 2. Horizontal divisions are each 200 Hz; vertical divisions are each 10 dB. Sampling bandwidth is 100 Hz. The center frequency is 4.000 MHz.

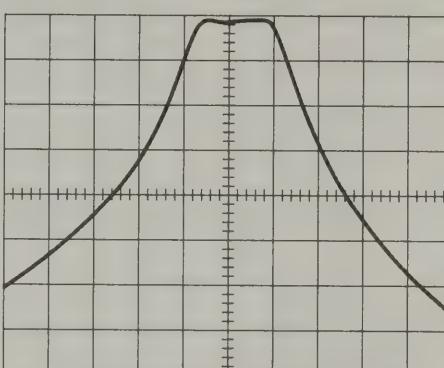


Fig E—Spectral photo showing the response of filter no. 3. Horizontal divisions are each 200 Hz; vertical divisions are each 10 dB. Sampling bandwidth is 100 Hz. The center frequency is 4.000 MHz.

Fig F—Spectral photo showing the response of filter no. 4. Horizontal divisions are each 200 Hz; vertical divisions are each 10 dB. Sampling bandwidth is 100 Hz. The center frequency is 4.000 MHz.

with the termination resistors, but insertion loss increases significantly. For CW use, however, this filter is probably more than adequate.

Filter no. 3 shows a bandwidth of approximately 350 Hz at the -3 dB points. It is symmetric and shows low ripple. This is a very good CW filter.

Filter no. 4 is a good example of what can be accomplished with high-quality crystals and proper terminations. This filter is used in a CW receiver designed by Dave Newkirk, AK7M (see cover of this issue). The input and output impedances of this filter are 200 ohms. To match the 50-ohm impedance of the test setup, 4:1 transformers were used. Filter insertion loss is 2 dB, with an ultimate rejection of over 90 dB.

The SSB filter shows a -3 dB bandwidth of approximately 2.1 kHz. There is no discernible ripple, and the insertion loss is 4.4 dB.

Conclusions

The empiric approach to designing Cohn filters for CW or SSB use is a viable alternative to purchasing commercial filters. The relatively high component cost for the best filter design tested (CW filter no. 4) still

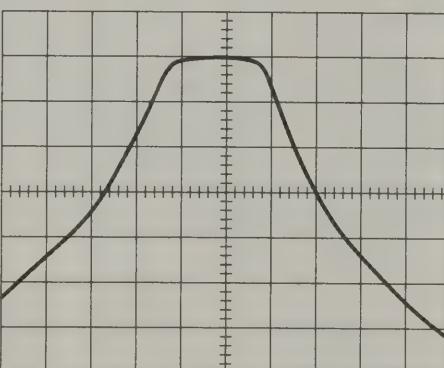


Fig G—Spectral photo showing the response of the SSB filter. Horizontal divisions are each 1 kHz; vertical divisions are each 10 dB. Sampling bandwidth is 100 Hz. The center frequency is 12.000 MHz.

results in an advantage of over 50% when compared to the price of commercial equivalents. All of the filters tested are adequate for most home-brew projects. They are fun to build, and result in appreciable savings.—Bruce O. Williams, WA6IVC, ARRL Staff

computer was used in the "construction" of a filter with six crystals. The circuit, again a narrow CW filter, is shown in Fig 5. The 200-pF capacitors used in the earlier filter are retained. The frequency response of this six-crystal filter is shown in Fig 6, where the "reference sweep" is the response of the previous three-element filter. The new filter has a -3 dB bandwidth of 354 Hz, but much steeper skirts than the three-element filter.

A Simple SSB Filter

Table 1 shows a number of simple three-pole filter configurations. Bandwidth is increased for a given set of crystals merely by decreasing the capacitance value. The frequency domain response for a three-pole SSB filter with 30-pF capacitors is shown in Fig 7. The "reference sweep" is the response of the earlier three-pole CW filter with 200-pF capacitors. The skirt response of the SSB three-crystal filter is certainly less than spectacular. More crystals will improve this response significantly. This simple three-pole filter is still practical for some applications, however, such as a portable VHF SSB transceiver.

Table 1
Cohn Three-Crystal Filter

Bandwidth (Hz @ -3 dB)	C (pF)	R_{end} (Ohms)
380	200	150
600	130	238
1.0k	70	431
1.8k	30	1.5k
2.5k	17	3.3k

Experimental Methods

The computer-based "experiments" have proved to be useful. There are generally no surprises. I've "built" filters on the computer using more than a dozen crystals. Some of the more practical designs have been transferred to hardware for receiver applications. Many of these designs operate at different frequencies, some using 4.433-MHz European TV color-burst crystals. These crystals are harder to obtain, but their frequency is more compatible with the existing HF ham bands, avoiding the spurious responses that can sometimes occur with a 3.579-MHz IF.

Almost all of my test equipment is built for an input and/or output impedance of 50 ohms. The test equipment is still easily used for filter experiments. Extra resistance is merely added at the filter input and output to bring the level up to that desired. This is illustrated in Fig 8. Ferrite transformers may also be built to transform impedance levels, but they cannot be changed as quickly as resistors.

It is often convenient to experiment with a filter that is contained within a receiver or transmitter. An example is shown in the

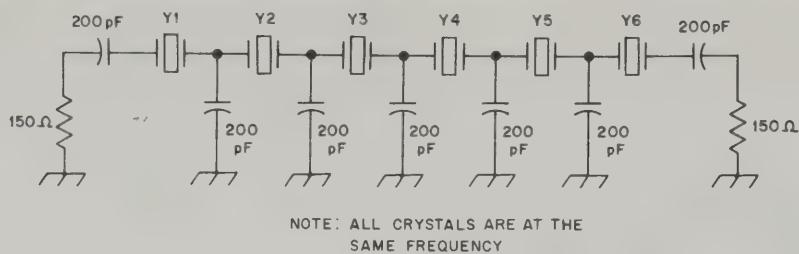


Fig 5—Circuit of a Cohn filter using six crystals.

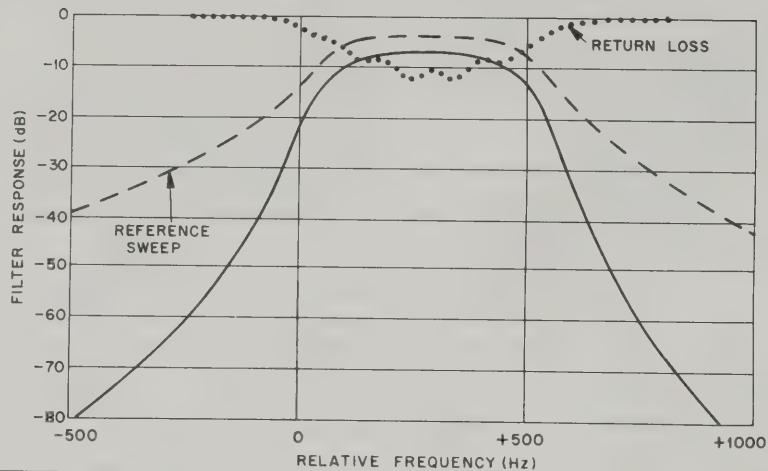


Fig 6—Frequency response of the six-crystal filter. The reference sweep is the response of the three-crystal filter of Fig 2.

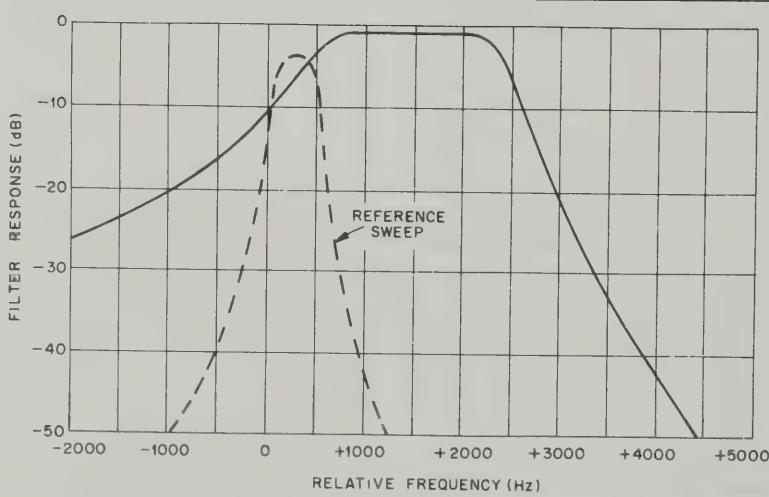
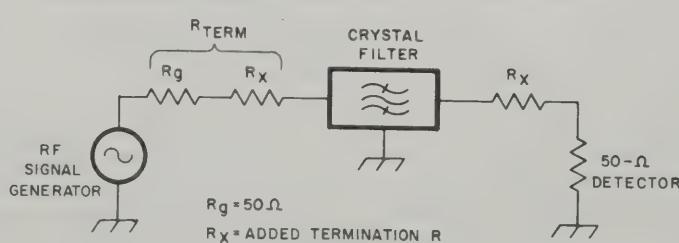


Fig 7—Frequency response of a simple three-pole SSB filter. The circuit is that of Fig 2 with all capacitors changed to 30 pF and terminations of 1500 ohms.



partial schematic of Fig 9A. Q1 is a dual-gate MOSFET mixer. The drain resistor determines the input loading impedance for the filter. An identical resistor terminates the filter output. An NPN amplifier, Q2, buffers the output—insurance that the following stages will not alter the crystal filter termination. Fig 9B is a modified form of the same filter. Tuned circuits have been inserted to present higher impedances to the transistors, affording more gain. The output amplifier is changed to a JFET. This modified circuit is better suited to higher impedance filters, as might be encountered with an SSB transmitter or receiver. Once the circuit containing the filter is built, filter response may be measured by tuning the receiver through a steady carrier while observing the output of a later stage with an oscilloscope or RF voltmeter.

It's often difficult to build a filter while also building a receiver. If problems occur, it is hard to tell if they are related to the filter or to the rest of the circuitry. Uncertainty is removed if receiver construction begins with a simpler, single-crystal filter. This allows you to get the receiver working before pursuing the better filter. I don't encourage you to retain the single-crystal filter as a final option. The enhanced performance afforded by additional crystals is more than ample justification for the minimal added effort and expense.

Other Crystals

The examples presented have used readily available color-burst crystals. There is nothing special about them. Indeed, they often represent the poorest possible quality for a crystal, and their frequency (3.579 MHz) can cause compatibility problems in many of the ham bands. They are, however, both available and cheap.

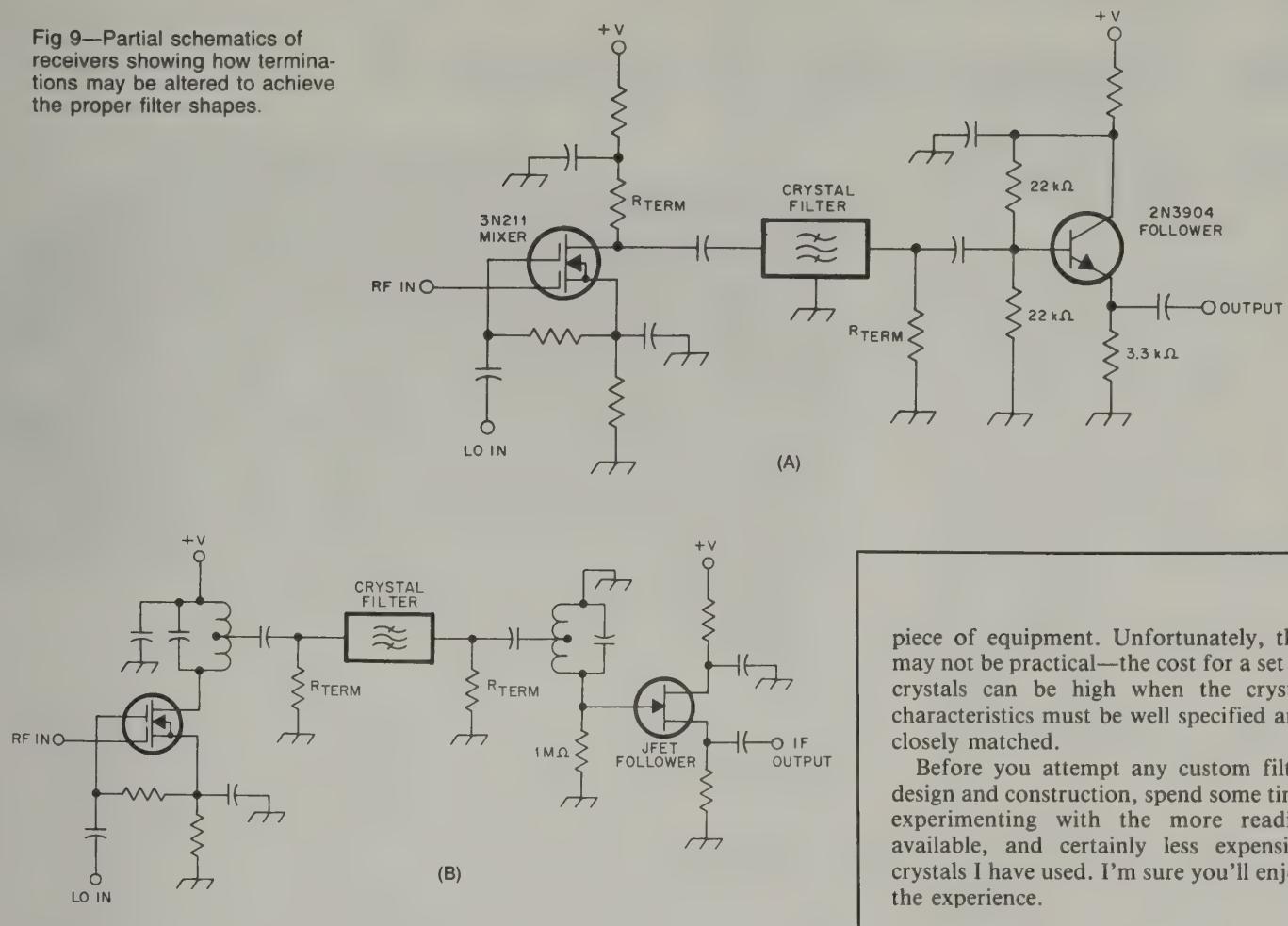
Many parts distributors list crystals for microprocessor applications in their catalogs. The only experience I have had with these crystals was with two 4-MHz crystals.⁶ The average Q was 150,000, motional inductance was 148 mH and the two crystals differed in frequency by 105 Hz. Further data on other crystal types would be of great use to the amateur community. Anyone out there with data to share? [See the sidebar to this article.—Ed.]

Traditional intuition might suggest that narrow-bandwidth filters are more difficult to design and build than those with wider bandwidth. Just the opposite is true; CW filters are easier to build than SSB or AM filters. This is fortunate, for it seems that much of the present home-brew activity is aimed at CW rigs.

Narrow-bandwidth CW filters are easily built with the lower frequency crystals, such

Fig 8—External resistors may be added to an experimental filter to allow use of 50-ohm instrumentation for circuit evaluation.

Fig 9—Partial schematics of receivers showing how terminations may be altered to achieve the proper filter shapes.



as those at 3.579 MHz. While an SSB filter can be built at 3.579 MHz, probably higher terminating impedances will be required. The termination value drops with increasing frequency, making wider bandwidth filters more easily realized at higher frequencies. I often build equipment with a 10-MHz IF because crystals with excellent Q are readily available for this frequency.

Typical parameters for these crystals are: motional inductance = 20 mH, parallel C = 3 pF and Q = 200,000. These characteristics result in practical CW filters with terminating impedances as low as 50 ohms, and SSB filters with 200- to 500-ohm loads.

You can, of course, order high-quality crystals for any desired frequency. It is then possible to fit a new filter into an existing

piece of equipment. Unfortunately, this may not be practical—the cost for a set of crystals can be high when the crystal characteristics must be well specified and closely matched.

Before you attempt any custom filter design and construction, spend some time experimenting with the more readily available, and certainly less expensive crystals I have used. I'm sure you'll enjoy the experience.

Notes

¹W. Hayward, "A Unified Approach to the Design of Crystal Ladder Filters," *QST*, May 1982, pp 21-27.

²S. Cohn, "Dissipation Loss in Multiple Coupled Resonators," *Proceedings IRE*, Aug 1959.

³W. Sabin, "The Solid-State Receiver," *QST*, Jul 1970, p 35.

⁴D. DeMaw, "His Eminence—The Receiver," *QST*, Jun and Jul 1976.

⁵A. I. Zverev, *Handbook of Filter Synthesis* (New York: John Wiley and Sons, 1967).

⁶Mouser Electronics, 11511 Woodside Ave, Lakeside, CA 92040, part no. ME332-1040.

A Tester for Crystal F, Q and R

This instrument allows you to analyze the crystals you want to use in IF ladder filters. The design is based on Wes Hayward's circuits in his classic *QST* ladder-filter paper.

By Doug DeMaw, W1FB

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Do you need a CW or SSB filter for a new project? Does the cost of commercial filters curl your hair? You can save dollars and design a filter for your needs by building this tester and following the instructions in Wes Hayward's *QST* articles on this subject.¹

He described some circuits for crystal analysis in his 1982 article, but no PC-board patterns were offered and the practical data was not complete. I built two units that contained the circuits Hayward recommended, but I found them awkward to use because they weren't assembled for convenience. This third model puts all of the key circuits on one small PC board.² I incorporated some circuit changes and added some features you should find useful.

Page space does not permit repeating the high points of the W7ZOI *QST* articles. Emphasis is on how to build and use this instrument. There is nothing tricky about the construction, and the cost is modest, especially if you have a reasonable supply of parts on hand.

Circuit Notes

Fig 1 shows the circuit for the crystal evaluator. A 20-dB, 50- Ω attenuator is used at the input port (J1). This helps to prevent the signal generator from overloading the broadband amplifiers, Q1 through Q4. It also provides a 50- Ω input termination for the crystal under test. These considerations are important if we are to obtain accurate



indications of the crystal characteristics. Q1 provides a 50- Ω load for the output of the crystal.

I added S1 to permit switching between the crystal and R4. Originally, I had to plug the potentiometer into the crystal socket from outside the panel. TP1 and TP2 are in parallel with R4 to allow resistance measurements from the front panel of the box. It is vital that the leads from the PC board to the Y1 socket, and from the board to R4, be kept very short. This minimizes stray inductance, which can cause improper readings, especially with crystals above 4 MHz.

Q1, Q2, Q3 and Q4 operate as broadband amplifiers. Total circuit gain is approximately 40 dB. A 3-dB attenuator is located immediately after Q2. The attenuator is switched in and out of the circuit by means of S2. Hayward provided a 3-dB gain reduction by changing the Q2 emitter components with a switch. I discovered that this approach could be tricky, depending upon the particular 2N3904 used. Some tailoring of values was necessary to ensure a 3-dB change. The resistive attenuator at S2 provides a dependable 3-dB drop in signal.

You can use a scope to measure changes in circuit response. J2 is provided for this purpose. If you choose this option, you can eliminate Q4 and the meter circuit. If Q4 is dropped from the circuit, terminate C10 (Q4 side) with a 56- Ω resistor. You can attach a frequency counter to J2. Hayward connected his counter to J1, but I found that my counter was not sensitive enough to trigger at that circuit point. The counter works fine at J2 because of the amplification provided by Q1, Q2 and Q3.

Q4 functions as an amplifier for the

metering circuit. D1 rectifies the RF energy to cause deflection of the needle of M1. R32 is a meter-sensitivity control.

I included the test oscillator, Q5, on the PC board. It is used for measuring crystal frequency via J3. This is essential when selecting closely matched crystals for a filter. Q5 functions as a Pierce oscillator. The feedback components (C18, C19) are suitable for 1 to 20-MHz crystals.

Construction Information

Fig 2 shows the interior of the tester. The circuit is housed in a Ten-Tec TW-38 cabinet. It measures (HWD) 3-1/16 x 8-1/4 x 6-1/8 inches. There is a lot of unused space in the box. You may want to miniaturize your version of this tester by choosing a smaller enclosure. The power supply (see Fig 3) can be eliminated if you already have an outboard power source (12 V dc at 100 mA or greater).

The PC board is mounted on metal spacers near the front panel. It is situated for the shortest leads practicable between the board and the related panel hardware. RG-174 miniature coaxial cable is used between J1 and the board; likewise from J2 to the PC board. Ground the shield braid at each end of these cables.

I used a Simpson 100- μ A meter for M1 of Fig 1. This is a high-priced component! You can use a low-cost edgewise surplus meter for M1. Most of these units have a 200- μ A basic movement, and they are plenty sensitive enough for this circuit. Look for these S meters and battery indicators in the surplus catalogs. It is a simple matter to paste a homemade numbered meter scale over the original one. This will provide better reference marks when checking the -3-dB points of the crystal response.

¹Notes appear at end of article.

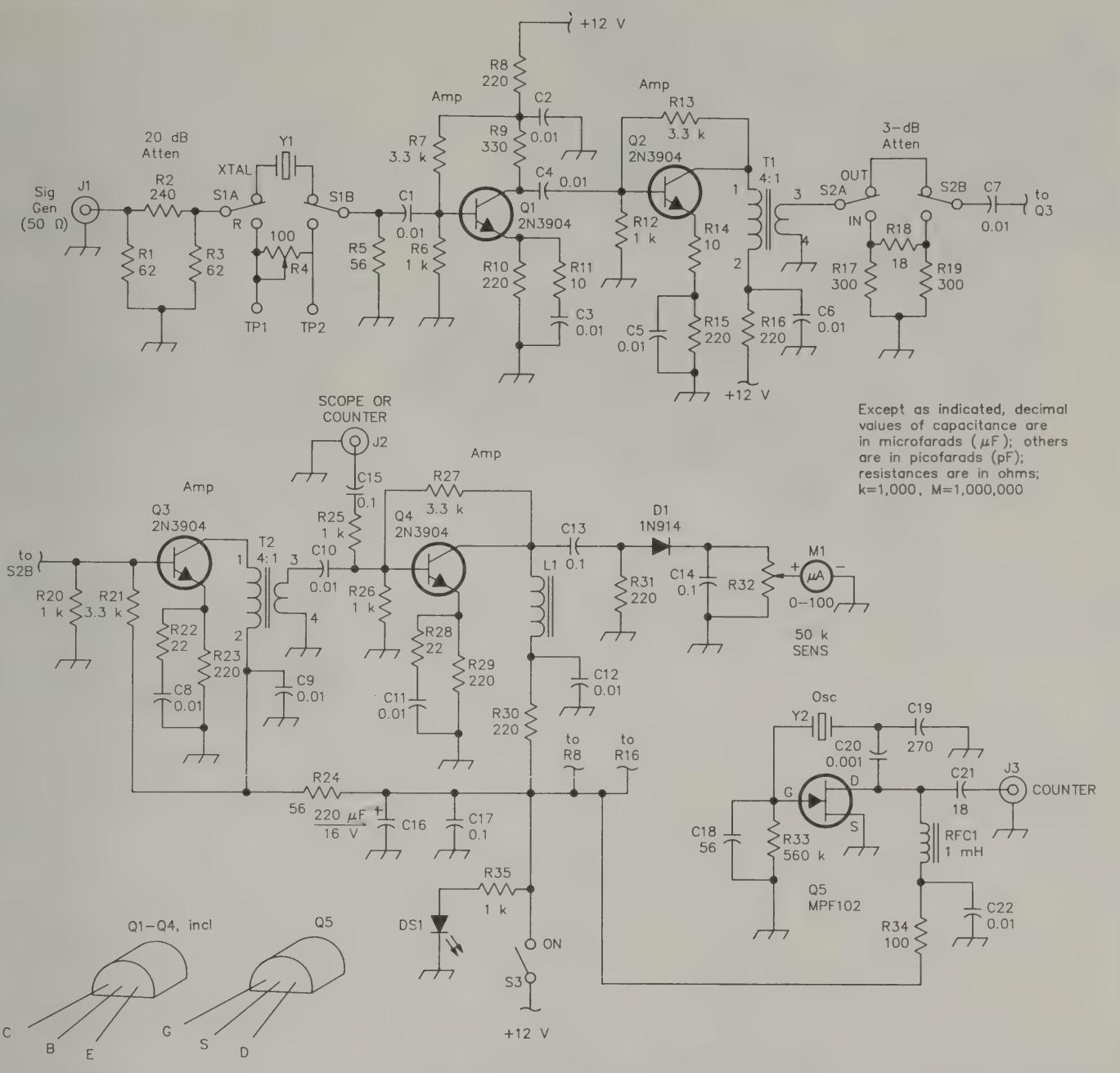


Fig 1—Schematic diagram of the RF portion of the crystal tester. Capacitors are disc ceramic, 100 V. Those with polarity marked are electrolytic or tantalum. Fixed-value resistors are 1/4-W carbon composition.

D1—Small-signal diode, type 1N914.

DS1—Red LED.

J1, J2, J3—BNC chassis-mount connector.

L1—RF choke. Use 15 turns of no. 26

enam wire on an Amidon Assoc FT-37-43 (850 μ) toroid.

M1—Small dc meter, 100 or 200 μ A (see text).

R4—100- Ω , linear-taper carbon-composition control. A 250- Ω control is also suitable.

R32—50-k Ω , linear-taper carbon-composition control.

RFC1—Miniature 1-mH RF choke.

S1, S2—DPDT toggle or slide switch.

T1, T2—4:1 broadband transformer. Use 12 primary turns of no. 26 enam wire on an

Amidon FT-37-43 toroid. Secondary has 6 turns of no. 26 enam wire.

TP1, TP2—Small press-in pin-type feed-through bushing or small binding post.

Y1, Y2—Crystal to be tested. Use crystal socket of choice or parallel sockets for types of holders desired.

T1 of Fig 3 is a surplus 12-V, 250-mA transformer. The output from the bridge rectifier, at the filter capacitor, is 16.9 V. A 12-V three-terminal regulator can be used in place of the LM317T variable regulator I use. Be sure to put a small heat sink on the voltage regulator, U2. My rectifier/regulator module is a unit (modified) I ob-

tained from a surplus dealer. There is no PC pattern for this assembly.

The Basics of Tester Use

Crystals used in ladder filters need to be matched within 100 Hz of one another—the closer the better, of course. Plug the crystal into the Y2 socket and attach a fre-

quency counter to J3. Select a set of crystals that is closely matched.

Next, insert one of the selected crystals into the Y1 socket. Connect a signal generator to J1 and place S1 in the XTAL position. Set S2 for 0 dB and set R32 at midrange. Carefully adjust the generator to the crystal frequency, as observed by a

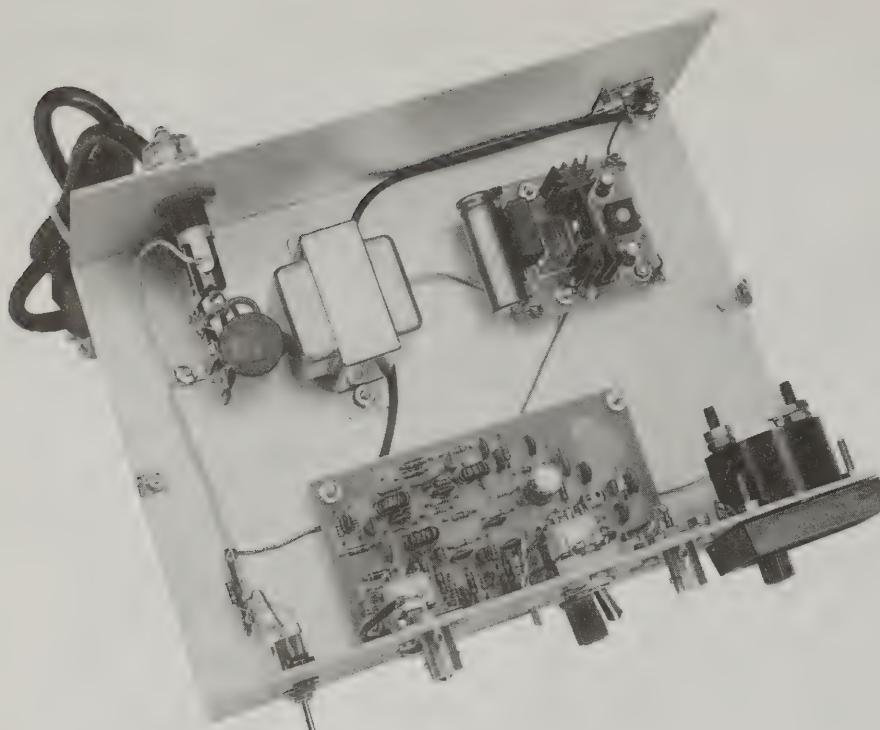


Fig 2—Interior view of the tester. Note the RF PC board close to the front panel to ensure short leads to the panel hardware. J2 is on the rear panel. Terminal strips are used as tie points for the ac line and DS1. R35 of Fig 1 is mounted on one of the terminal strips.

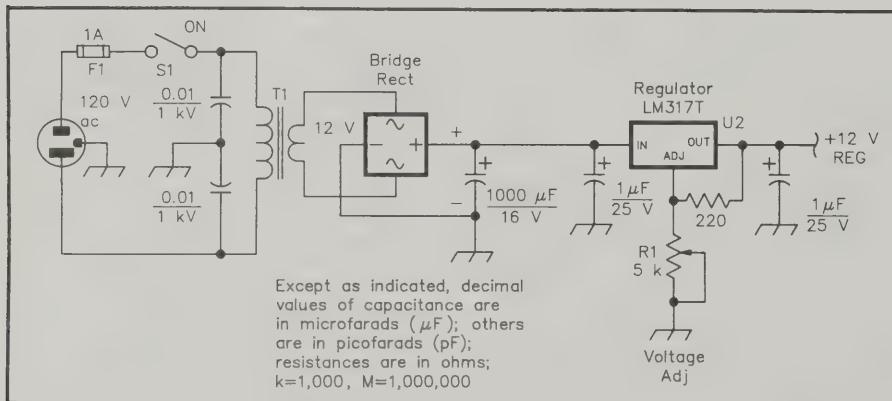


Fig 3—Circuit of the 12-V power supply. Decimal-value capacitors are disc ceramic, 1 kV; those with polarity marked are electrolytic or tantalum.

F1—1-A fuse in panel-mount holder.

R1—Linear taper PC-mount control, 5 kΩ.

S1—SPST toggle.

T1—12-V, 250-mA power transformer.

U1—Bridge rectifier, 50 PRV, 1 A.

U2—Adjustable voltage regulator or fixed-voltage unit (see text).

peak reading obtained on M1. This is a touchy procedure, since you can pass across the crystal series-resonant frequency without seeing the needle of M1 move! A generator with a vernier knob is essential for this test unless you are quite steady of hand. Note the frequency at which peak response occurs.

Now, switch in the 3-dB pad and log the meter reading. It will drop to the left. Remove the 3-dB pad and adjust the generator frequency above and below the crystal series-resonant frequency. Find the two

points where M1 reads the same as with the 3-dB pad in the line. These are the 3-dB response points for the crystal. Use your frequency counter to measure the frequency at the 3-dB points. With this information you can determine the crystal Q, in accordance with the Hayward article referenced in Note 1.

The last procedure calls for relocating the crystal series-resonant frequency (maximum M1 reading). Place S1 in the R position and adjust R4 to obtain the same meter reading that was noted with S1 in the XTAL

mode. Now, measure the resistance of R4 across TP1 and TP2. This provides the resistive characteristic of your crystal. This data is essential for designing your filter, as discussed in Hayward's article.

Use the lowest signal-generator output amplitude practicable, consistent with a midscale meter reading at crystal resonance. Feeding too much driving power into the tester will lead to inaccurate test results, and excessive drive can damage the crystal under test.

Some Final Remarks

I hope you will be encouraged to build your own crystal filters for transmitters and receivers. Certainly, the Hayward procedure is simple and easy to duplicate. The practical data for this tester is meant to stimulate you toward doing your own thing with crystal filters. Surplus computer crystals are widely available at low cost. Check the surplus catalogs for these units; most are in the \$1 class.

Notes

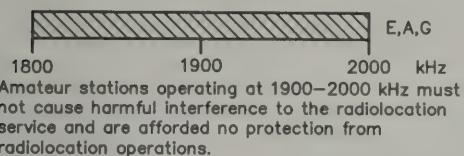
¹W. Hayward, W7ZOI, "A Unified Approach to the Design of Crystal Ladder Filters," *QST*, May 1982, p 21. Also see Hayward's "Designing and Building Simple Crystal Filters," *QST*, July 1987, p 24.

²Drilled and plated PC boards for this project are available from FAR Circuits (N9ATW), 18N640 Field Ct, Dundee, IL 60118.

US AMATEUR BANDS

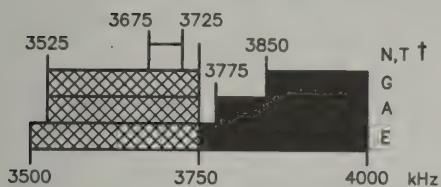
February 1, 1994

160 METERS



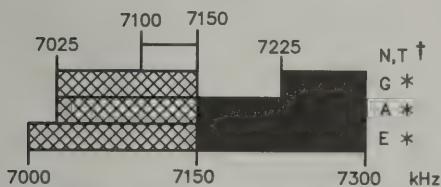
Amateur stations operating at 1900–2000 kHz must not cause harmful interference to the radiolocation service and are afforded no protection from radiolocation operations.

80 METERS



5167.5 kHz (SSB only): Alaska emergency use only.

40 METERS



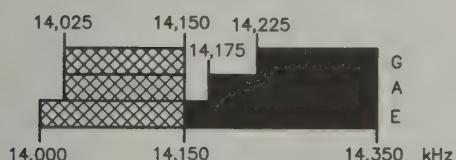
* Phone operation is allowed on 7075–7100 kHz in Puerto Rico, US Virgin Islands and areas of the Caribbean south of 20 degrees north latitude; and in Hawaii and areas near ITU Region 3, including Alaska.

30 METERS

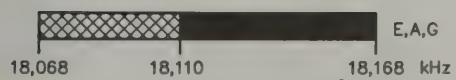


Maximum power on 30 meters is 200 watts PEP output. Amateurs must avoid interference to the fixed service outside the US.

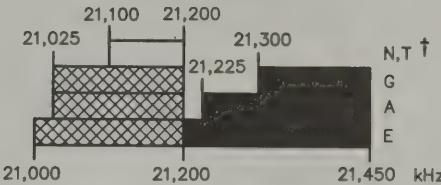
20 METERS



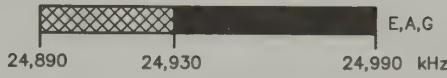
17 METERS



15 METERS



12 METERS



10 METERS



Novices and Technicians are limited to 200 watts PEP output on 10 meters.

6 METERS



2 METERS



1.25 METERS

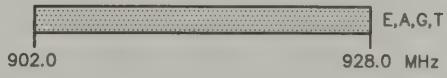


Novices are limited to 25 watts PEP output from 222 to 225 MHz.

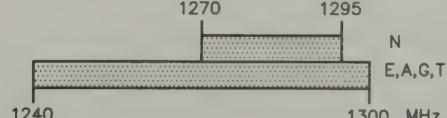
70 CENTIMETERS **



33 CENTIMETERS **



23 CENTIMETERS **



Novices are limited to 5 watts PEP output from 1270 to 1295 MHz.

US AMATEUR POWER LIMITS

At all times, transmitter power should be kept down to that necessary to carry out the desired communications. Power is rated in watts PEP output. Unless otherwise stated, the maximum power output is 1500 W. Power for all license classes is limited to 200 W in the 10,100–10,150 kHz band and in all Novice subbands below 28,100 kHz. Novices and Technicians are restricted to 200 W in the 28,100–28,500 kHz subbands. In addition, Novices are restricted to 25 W in the 222–225 MHz band and 5 W in the 1270–1295 MHz subband.

Operators with Technician class licenses and above may operate on all bands above 50 MHz. For more detailed information see The FCC Rule Book.

KEY

- diagonal lines = CW, RTTY and data
- diagonal lines with dots = CW, RTTY, data, MCW, test, phone and image
- solid black = CW, phone and image
- horizontal lines = CW and SSB
- diagonal lines with dots = CW, RTTY, data, phone, and image
- white = CW only
- E = EXTRA CLASS
- A = ADVANCED
- G = GENERAL
- T = TECHNICIAN
- N = NOVICE

† Only Technician-class licensees who have passed a 5-WPM code test may use these frequencies.

** Geographical and power restrictions apply to these bands. See The FCC Rule Book for more information about your area.

Above 23 Centimeters:

All licensees except Novices are authorized all modes on the following frequencies:

2300–2310 MHz
2390–2450 MHz
3300–3500 MHz
5650–5925 MHz
10.0–10.5 GHz
24.0–24.25 GHz
47.0–47.2 GHz
75.5–81.0 GHz
119.98–120.02 GHz
142–149 GHz
241–250 GHz
All above 300 GHz



For band plans and sharing arrangements, see The ARRL Operating Manual or The FCC Rule Book.



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- The AC/RF Sources chapter explains, clearly and concisely, modern oscillator and synthesizer design.
- Mathematics for Amateur Radio provides a refresher course in the math concepts that form the basis of ham radio and electronics.
- The Station Setup and Accessories chapter shows how to set up or enhance your station for top performance; among other projects, it includes a boom/headset mike and three computer interfaces.
- In the Transceivers chapter, you'll learn about the systems and design that go into modern radio equipment. Projects include a beginner's shortwave receiver, QRP transceivers and a potent 50-W solid-state linear power amplifier.
- The References chapter organizes, in one convenient place, the tables and charts hams need and use most often.

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The Radio Amateur is:

CONSIDERATE . . . never knowingly operates in such a way as to lessen the pleasure of others.

LOYAL . . . offers loyalty, encouragement and support to other amateurs, local clubs, and the American Radio Relay League, through which Amateur Radio in the United States is represented nationally and internationally.

PROGRESSIVE . . . with knowledge abreast of science, a well-built and efficient station and operation above reproach.

FRIENDLY . . . slow and patient operating when requested; friendly advice and counsel to the beginner; kindly assistance, cooperation and consideration for the interests of others. These are the hallmarks of the amateur spirit.

BALANCED . . . radio is an avocation, never interfering with duties owed to family, job, school, or community.

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—The original Amateur's Code was written by Paul M. Segal, W9EEA, in 1928.—



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About the Author

M. F. "Doug" DeMaw, W1FB was licensed in 1950 as W8HHS while he worked as an R&D engineer for the University of Michigan Willow Run Research Center. Later, he worked as an R&D engineer for Ryan Aeronautical Research Laboratories in San Diego. He served also as a TV broadcast engineer at WWTW, channel 9, and as chief engineer/DJ for WATT radio in Cadillac, Michigan.

Doug founded Avtronics, Inc in 1960 at Traverse City, Michigan, where his firm manufactured low-frequency radio-beacon transmitters and alarm receivers for civilian airports. Upon selling Avtronics in 1963 he established Comaire Electronics in Ellsworth, Michigan, where the firm manufactured VHF and UHF amateur equipment, plus low-frequency radio-beacon transmitters and VHF Unicom transceivers. While operating Avtronics, he established *VHFer* magazine, which he turned over to Loren Parks, K7AAD, in 1965 when he joined the ARRL HQ staff as an assistant technical editor.

In 1968 he was promoted to *Handbook* Editor/Lab Supervisor, and in 1970 he succeeded the late George Grammer, W1DF, as Technical Department Manager/Senior Technical Editor.

While employed by the ARRL, he coauthored *Solid State Design for the Radio Amateur* with well-known QRPer Wes Hayward, W7ZOI. W1FB also wrote two books for Prentice-Hall, Inc and one book for Howard Sams.

W1FB retired early from the ARRL staff in 1983 to return to his native state of Michigan. Since retiring, he has written articles for *QST* and has done article and book editing for the ARRL.

Doug is a Life Member of the ARRL, a past Senior Member of the IEEE and past president of the Michigan QCWA, Chapter 10. He holds a Radiotelephone First Class license and an Amateur Extra Class license. He is presently Chairman of the Lake County Board of Commissioners in Michigan and is a member of the Michigan Association of Counties.

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